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GURBATOV, S. N., Gorkiy State University

[Abstract] A study is made of the parametric interaction in dispersionless media when the parameter wave or pumping wave has the same nature as the signal. The pumping wave can be excited either at the boundary or within the medium by a distributed synchronous source. Based on the heterogeneous Burgers equation, the specifics of transformation of time and energy characteristics of nonlinear waves interacting with the leading or trailing edges of pumping pulses excited by external wideband distributed sources are studied. It is shown that in this case the heterogeneous Burgers equation for replacement of variables is reduced to a homogeneous equation, and the field of the signal interacting with the pumping wave is similar to a field in free space in a certain new effective medium cross section. The author thanks A. N. Malakhov and A. I. Saichev for helpful comments.
Figures 4; references 23: 17 Russian, 6 Western.
[240-6508]

ANTENNAS AND PROPAGATION

UDC 621.396.96:621.391.2

INCOHERENT SERIAL SIGNAL DETECTION AGAINST BACKGROUND OF PASSIVE INTERFERENCE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian
Vol 26, No 4, Apr 83 (manuscript received 12 Jan 82) pp 58-62

[Article by A.M. Shloma and G.B. Gol'fel'd]

[Text] Aspects of nonlinear autoregression transformation are examined. Algorithms for incoherent serial signal detection against the background of passive interference are synthesized, and effectiveness estimates are obtained.

Coherent serial signal detection against the background of passive (dependent stationary) interference was examined in [1,2] under conditions of linear conversion of the observed process on the basis of the autoregression approach. However, this process often undergoes nonlinear conversion in practice. The development of autoregression sequential analysis methods for this case is of interest. Two aspects of this problem are investigated below. In the first, the observed process is continuous; however, when digitized in time the samples are correlated. In the second, the processes are discrete.

Continuous observed process. Let there be a continuous process which is time-digitized during processing and for which the samples are correlated.

Let said processing be nonlinear inertialess conversion. The set of samples of the observed process $\{x_i | i=\overline{0,n}\}$ prior to transformation permits autoregression representation.

We shall now find the autoregression representation of the process after nonlinear transformation. In order to do this, we employ a general autoregression equation of order l, reduced to first-order vector equation [3]

$$\tilde{\vec{x}}_i = \tilde{\mathbf{P}}\tilde{\vec{x}}_{i-1} + \tilde{\vec{\varepsilon}}_i, \quad i=\overline{1,n}, \tag{1}$$

where $\tilde{\vec{x}}_i = \{x_{i-k} | k=\overline{0, l-1}\}$; $\tilde{\vec{\varepsilon}}_i = \{\varepsilon_{i-k} | k=\overline{0, l-1}\}$; $\varepsilon_{i-k}=0$ for $k\neq 0$; $\underline{\tilde{\mathbf{P}}} = \{\rho_{jk} | j, k=\overline{1, l}\}$ & $\rho_{1k}=\rho_k$, $\rho_{k+1,k}=1$, $\rho_{jk}=0$ otherwise.

Thus, (1) is a special case of the first-order vector stochastic difference equation

$$\vec{x}_i = \mathrm{P}\vec{x}_{i-1} + \vec{\varepsilon}_i, \tag{2}$$

where $\vec{x}_i = \{x_{i-k} \mid k = \overline{0, l-1}\}$, $\vec{x}_i \backsim N(0;\ \Sigma_x)$; $\vec{\varepsilon}_i = \{\varepsilon_{i-k} \mid k = \overline{0, l-1}\}$, $\vec{\varepsilon}_i \backsim$ $\backsim N(0;\Sigma)]$; $\mathrm{cov}\, \vec{\varepsilon}_i \vec{\varepsilon}_j{}' = 0$, if $i \neq j$.

Therefore, the discussion below will refer to (2). It is easy to find the connection between Σ_x and Σ: $\Sigma = \Sigma_x - \mathrm{P}\Sigma_x\mathrm{P}'$, $\mathrm{P} = (\rho_{ij})$ -- autocorrelation matrix of process (2). In accordance with [4] we switch from (2) to the stochastic differential equation in Ito form

$$\overrightarrow{dx(t)} = (R - I)\overrightarrow{x(t)}\, dt + \overrightarrow{du(t)}, \tag{3}$$

$\overrightarrow{du(t)} = \overrightarrow{\varepsilon(t)}\, dt$ — Wiener process differential,

$$\left.\begin{aligned} R &= \lim_{\Delta t} \mathrm{P}; \quad \overrightarrow{\varepsilon(t)} = \lim_{\Delta t} \vec{\varepsilon}_i, \quad \overrightarrow{\varepsilon(t)} \backsim N(0;\ \Psi) \\ \Psi &= \lim_{\Delta t} \Sigma; \quad \overrightarrow{x(t)} = \lim_{\Delta t} \vec{x}_i \end{aligned}\right\}, \tag{4}$$

$\Delta t = [i - (i-1)] \to 0$; $i \to t$; I — identity matrix.

The problem can now be formulated as follows. Let there be a stochastic Ito process $\overrightarrow{x(t)}$ described by (3). The inertialess nonlinear transformation $\overrightarrow{y(t)} = \phi[\overrightarrow{x(t)}]$ is performed over $\overrightarrow{x(t)}$. Find the Ito representation for the

process $\overrightarrow{y(t)}$.

In accordance with [4], $\overrightarrow{y(t)}$ is also an Ito process which satisfies the stochastic differential equation

$$\overrightarrow{dy(t)} = \left\{ \vec{x}'(t)(R'-I)\frac{d\varphi[\overrightarrow{x(t)}]}{\overrightarrow{dx(t)}} + \frac{1}{2}\operatorname{tr}\left[(\Psi^{\frac{1}{2}})' \frac{d^2\varphi[\overrightarrow{x(t)}]}{[\overrightarrow{dx(t)}]^2} (\Psi^{\frac{1}{2}}) \right] \right\} dt + \left[\frac{d\varphi[\overrightarrow{x(t)}]}{\overrightarrow{dx(t)}} \right]' \overrightarrow{du(t)}. \tag{5}$$

For quadratic transformation $\overrightarrow{y(t)} = [\overrightarrow{x(t)}]^2 = [x_1^2(t), \ldots, x_\ell^2(t)]'$ and (5) takes on the form

$$\overrightarrow{dy(t)} = [2\vec{x}'(t)(R'-I)\overrightarrow{x(t)} + \operatorname{tr}(\Psi^{\frac{1}{2}})'(\Psi^{\frac{1}{2}})]\,dt + 2\vec{x}'(t)\,\overrightarrow{du(t)}. \tag{6}$$

Passage to the limit, inverse to (4), makes it possible to obtain the stochastic difference equations corresponding to (5) and (6).

Nonlinear inertialess transformation of the autoregression process thus leads again to an autoregression-type process with parameters found from (5).

Discrete process. The initial data consists of the sample of observations $\{x_i | i = \overline{0,n}\}$ over the random quantity X, which is described by autoregression-type equation (1). We note that no previous investigation has been done on extending the temporal autoregression approach to the class of nonlinear observation models. While the situation has been formalized to some extent for continuous processes (even though the results obtained with the help of (5) do not always permit natural interpretation), no methods are known for synthesizing incoherent detection algorithms (either sequential or classical) for signals against the background of passive interference which allow for inter-period correlation of the interference. For example, the studies [5,6], although they note the occurrence of inter-period correlation of the passive interference with respect to envelope, no allowance is made for this in the synthesis.

A heuristic approach is proposed: the operation of nonlinear inertialess transformation $\{y_i\}=\phi[\{x_i\}]$ is reduced to independent transformation of each element of sample $\{x_i\}$

$$\{y_i\}=\{\varphi(x_i)\}. \tag{7}$$

The connection between the elements y_i is described by an equation of the type (1). The problem is to select the parameters of this equation in a manner which is acceptable for each specific case.

We shall consider the use of the proposed approach for synthesizing algorithms for incoherent sequential detection of quasi-determinate signals against the background of dependent stationary interference, which can be used in systems for selecting moving-target signals against the background of externally-coherent passive interference. The sign that a signal is present in these systems consists of amplitude fluctuations (nonstationarity in the average) of the observed process which are extracted by an amplitude detector (with a quadratic response for weak signals) which follows the intermediate-frequency amplifier. Since the signal in these systems is a sequence of pulses, i.e., the observed process is essentially discrete, we use representation (7) to solve the synthesis problem.

We shall be dealing below, by analogy with the method in [2], with a scalar Markov process, which is employed most often in practice. The initial statistical detection problem is described by a system of alternative autoregression equations for the hypotheses $H_0[x \sim N(0;\ \sigma_x^2]$ and $H_1[X \sim N(\mu;\ \sigma_x^2)]$:

$$\left.\begin{aligned} &H_0: x_i=\rho x_{i-1}+\varepsilon_i \\ &H_1: x_i-\mu\cos\psi_i=\rho\,(x_{i-1}-\mu\cos\psi_{i-1})+\varepsilon_i \end{aligned}\right\}, \tag{8}$$

where ψ -- Doppler phase of observation due to target movements. The transformation of process $\{x_i\}$ appears as $\{y_i\} = \{x_i^2\}$. Applying this to (8), and averaging with respect to the parameter ψ, which because of the incoherence of the problem is random and uniformly distributed over the envelope $(0,2\pi)$ and normalizing with respect to σ_x^2, we can write:

$$\left.\begin{aligned} H_0 &: t_i^2 = \gamma t_{i-1}^2 + \xi \varkappa_i^2 \\ H_1 &: t_i^2 \mp \delta^2/2 = \gamma\,(t_{i-1}^2 \pm \delta^2/2) + \xi \varkappa_i^2 \end{aligned}\right\}, \tag{9}$$

where $t_i^2 = x_i^2/\sigma_x^2,\ \varkappa_i^2 = \varepsilon_i^2/\sigma^2,\ \delta^2 = \mu^2/\sigma_x^2,\ \gamma = \rho^2,\ \xi = 1-\rho^2,\ \sigma^2 = \sigma_x^2(1-\rho^2).$

In accordance with [7], for H_0: $t_i^2 \sim \chi^2(1)$, $\kappa_i^2 \sim \chi^2(1)$; for H_1: $t_i^2 \sim \chi^2(\delta^2;1)$, $\kappa_i^2 \sim \chi^2(1)$. Considering this, it seems natural in (9) to center the non-centrality parameter δ^2 for H_1. In addition, since from the electrical engineering viewpoint the purpose of quadratic transformation is to ensure an unbiased criterion in order to make incoherent processing possible, the combination terms are disregarded when we go to (9). This fact, along with the averaging over ψ, results in losses which are typical when we go from coherent to incoherent processing. Modeling has shown that centering with different polarity makes it possible to equalize series (9) such that $\kappa_i^2(H_0) \approx \kappa_i^2(H_1)$ for the strong correlation of observations ($|\rho| \to 1$) and low signal/noise ratio ($\delta^2 \to 0$), when the interference envelope is a monotonic function, which is typical in moving-target radar. Here the upper combination of signs corresponds to an increase ($t_i^2 > t_{i-1}^2$), while the lower combination corresponds to a decrease ($t_i^2 < t_{i-1}^2$). We note that system (9), at least externally, is adequate to the initial model (8).

Synthesis of detection algorithms. After the corresponding conditional distributions have been obtained from (9), the sequential analysis procedure can be written as follows:

$$\left.\begin{aligned} & b \lesseqgtr \lambda_n \approx \sum_{i=1}^{n} z_i \lesseqgtr a \\ & z_i = -\frac{1}{2}\ln\left[1 \mp \frac{\delta^2(1+\gamma)/2}{t_i^2 - \gamma t_{i-1}^2}\right] \pm \frac{\delta^2}{4}\,\frac{1+\gamma}{\xi} \end{aligned}\right\}, \tag{10}$$

where a and b -- Wald thresholds. Expression (10) is difficult for implementation as well as analysis. In order to simplify procedure (10) we shall exploit the fact that the incoherent processing channel is usually a "standard electrical engineering section" consisting of an IF amplifier, a square-law detector and a low-pass filter in series. In accordance with [8], when the passband is small $\vartheta = \Delta_{LPF}/(0.5\Delta_{IFA})$ the process at the output of the section is normalized.

Considering normalization in (9) $\kappa_i^2 \sim N(1; \vartheta)$, and the sequential procedure takes on the form:

$$\left.\begin{aligned} & b \lesssim \lambda_n \approx \sum_{i=1}^{n} z_i \lesssim a \\ & z_i = \pm \frac{\delta^2 (1+\gamma)}{4\vartheta\xi^2}\left[2\,(t_i^2 - \gamma t_{i-1}^2) \mp \delta^2 (1+\gamma)/2 - 2\xi\right] \end{aligned}\right\} . \tag{11}$$

It is easy to see that (11) assumes weighted accumulation of the post-detector alternate-period differences.

Analysis of effectiveness of incoherent sequential detection algorithms. Considering the properties of model(9), it is easy to obtain the average time of procedure (11):

$$E(n \mid H_0) \approx \frac{(1-\alpha)\,b + \alpha a}{-\,\delta^4 (1+\gamma)^2/8\vartheta\xi^2}, \qquad E(n \mid H_1) \approx \frac{(1-\beta)\,a + \beta b}{\delta^4 (1+\gamma)^2/8\vartheta\xi^2}, \tag{12}$$

where α and β -- false alarm and missed-target probabilities.

By comparing (12) with the expressions obtained in [9] for independent observations, we can conclude that allowance for correlation of the observations in incoherent processing also makes it possible to reduce the average time required for the procedure. For example, it follows from (12) that as $\rho \to 1 E(n|\cdot) \to 0$, i.e., like for coherent processing, we arrive at singularity. Nonetheless, comparing (12) with the expression cited in [1] for coherent processing, we can state that for small signal/noise ratios ($\delta^2 \to 0$), like in classical analysis, incoherent detection is significantly poorer in terms of noise tolerance than coherent (this is expressed here in a significant increase in the number of observations required to make a decision).

As it turns out, if expression (6) is applied to model (8), it is transformed to a form analogous to (9) with parameters $\gamma=2\rho-1$ and $\xi=2(1-\rho^2)$. The form of notation for procedures (10) and (11), as well as the expressions for average analysis time (12) remain the same. By substituting new γ and ξ in (12), we are easily convinced that this approach leads to a less effective procedure than the heuristic approach. This is also confirmed by the modeling results.

Using the Stratonovich approach instead of the Ito approach (or, which is equivalent, ordinary analysis rules) leads, to within determinate constants, the same equations of system (9); therefore, the effectiveness of the corresponding procedures is also the same.

In conclusion, let us consider the effectiveness of sequential procedure (11) in relation to the corresponding classical procedure with a sample size N. We shall use as the effectiveness measure these quantities [10]: $\xi_0=[N/E(n|H_0)]-1$, $\xi_1=[N/E(n|H_1)]-1$. Since the classical procedure is based on the same likelihood ratio as the sequential procedure (11), by applying the method from [10] it can be demonstrated ξ_0 and ξ_1 in this case are determined by the expressions ([10], (5), (6), p. 162) for the case of coherent sequential detection when the observations are independent and identically distributed. This is explained by the fact that procedure (11) is based on accumulating alternate-period differences, which in accordance with the present autoregression approach are independent.

It follows from analyzing the functions $\xi_0(\alpha)$ and $\xi_1(\beta)$ that the use of sequential analysis provides a significant increase in the effectiveness of systems for selecting moving targets against the background of externally-coherent passive interference, especially for $\alpha<<\beta$, when $\alpha\to0$, β=const $\xi_0\to\infty$.

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ANALYSIS OF EFFECTIVENESS OF SPACE-TIME RADIO SIGNAL PROCESSING IN PRESENCE OF INTERFERENCE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 4, Apr 83 (manuscript received 18 Jun 82 after revision) pp 54-58

[Article by A.S. Popov, V.V. Davydenko and Yu.V. Skobel'tsev]

[Text] This article examines the problem of optimizing the weight coefficient vector in an adaptive antenna array. The array functioning algorithm is synthesized by the variable-state method, based on Kalman filtering. The Results of calculating the effectiveness of space-time processing algorithms for a flat equidistant adaptive array with different numbers of receiving elements, noise power and variation in the angle of arrival of one and two interference events are presented.

The solution to the problem of space-time processing of signals and noise consists of determining the optimal value of the weight coefficient vector (WCV) in real time, which is equivalent to synthesizing the optimal directivity pattern (DP) of the active antenna array (AAA) [1]. Various methods and criteria can be used to determine the optimal value of the WCV. One of the most applicable criteria in developing AAA for a communications system system, which allows constructive solutions to be obtained, is the criterion of the minimum mean square deviation of the received signal from a standard signal (MMSD) [1-3]. For the algorithms which operate in accordance with this criterion developed by Widrow [2] the optimal WCV value of $W^T = [w_1, w_2, \ldots, w_n]$ can be found by solving the Wiener-Hopf vector matrix equation

$$W = R^{-1}B, \tag{1}$$

where $R=E[X^*, X^T]$ covariation matrix of vector of received signals $X^T = [x_1, x_2, \ldots, x_n]$; $B=E[X^*, g]$ -- covariation between vector X and desired

signal g; E -- sign of mathematical expectancy; * -- symbol for complex conjugation. The value of the WCV can be obtained recursively by solving the Widrow-Hoff equation [2]

$$W(k+1) = W(k) + 2\mu [B - RW(k)] \quad (2)$$

or the corresponding differential equation

$$\frac{dW(t)}{dt} = 2\mu\varepsilon(t)X(t), \quad (3)$$

where $\varepsilon(t)$ -- error signal, defined as the difference between the received signal $W^T X$ and standard g; μ -- coefficient determining stability and rate of convergence of algorithm.

When the coefficients μ in equations (2) and (3) satisfy the Dvoretskiy conditions these algorithms represent a stochastic approximation procedure of the Robbins-Monroe type [4]. The solution obtained through their use converges to a constant WCV value: $W(t) \to const$ as $t \to \infty$. In addition, it turns out in many practical cases that the signal-interference situation fluctuates continuously because of changes in propagation conditions or other factors; therefore, procedures (2) and (3) are non-optimal.

With randomly fluctuating space-time parameters of the signals and interference it is best to use the estimate of the WCV based on variable-state methods [5], which admits allowance for the occurrence of perturbations. In order to do this, the process by which the optimal values of the weight coefficients varies is represented in the form of the stochastic differential state equation

$$\frac{dW(t)}{dt} = F(t)W(t) + G(t)U(t), \quad (4)$$

where W(t) -- value of WCV determining the state; F(t) and G(t) -- nxn state matrix and excitation matrix, respectively; U(t) -- generating Gaussian n-dimensional white noise with null mean and covariation matrix $E[U(t)U(\tau)] = Q_U(t)\delta(t-\tau$; $Q_U(t)$ -- noise intensity; $\delta(t-\tau)$ -- Kronecker's symbol.

The observation equation can be represented as

$$Z(t) = H(t)W(t) + V(t), \quad (5)$$

where $Z(t)$ -- signal at output of common adder; $H(t)=X^T(t)$ -- observation matrix, consisting of the combination of the valid signal and interference at the output of the AAA; $V(t)$ -- observation noise with statistical characteristics $E[V(t)]=0$ and $E[V(t)V(\tau)]=Q_V(t)\delta(t-\tau)$.

The characteristic difference between the approach used here and the traditional Kalman approach is that the state vector $W(t)$ in equation (4) is not the valid received signal; the latter is contained in observation matrix $H(t)$. In other words, it is not the received realization which is estimated in this case, but rather its linearly connected weight vector-function $W(t)$.

The rule for obtaining the estimate of the WCV which is optimal with respect to the selected criterion is expressed in the form of a stochastic differential estimation equation [5]

$$\frac{d\hat{W}(t)}{dt} = F(t)\hat{W}(t) + K(t)[Z_э(t) - H(t)\hat{W}(t)]. \qquad (6)$$

Here $K(t)$ -- gain of Kalman filter, found from the formula

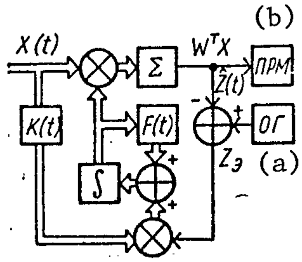
$$K(t) = D(t)H^T(t)Q_V^{-1}(t).$$

The principle by which the a posterior error dispersion $D(t)$ fluctuates over time is described by the Riccati matrix equation

$$\frac{dD(t)}{dt} = F(t)D(t) + D(t)F^T(t) - D(t)H^T(t)Q_V^{-1}(t)H(t)D(t) + \\ + G(t)Q_U(t)G^T(t).$$

As can be seen from the expressions above, one characteristic of this WCV estimation procedure is the fact that, regardless of the linearity of the estimation problem with respect to state (4) and observation (5), the value of the a posteriori dispersion depends upon the observation results, which requires that the corresponding equation be solved in real time. Figure 1 shows the structural diagram of a device which operates in accordance with algorithm (6). In the diagram (a) -- reference oscillator, (b) -- receiver. Specially generated signals or sync signals can be used as the standard Z_e. In some cases the standard can be generated from the information signal.

Fig. 1



Obviously, when $F(t)\equiv 0$ and $G(t)\equiv 0$ WCV estimation algorithm (6) takes on the form (2) or (3), and can be viewed as a more general algorithm which can be employed in more complex statistical signal-interference situations. This algorithm can be extended further for nonlinear and nonstationary conditions by using standard methods [5].

We shall now examine the effectiveness of the algorithms for optimal WCV estimation in different statistical situations. The indicator of the effectiveness of space-time signal and interference processing algorithms can be the coefficient of variation of the signal/(interference+noise) levels at the output of the processing device with respect to the input

$$\eta=[P_{s\,out}/(P_{i\Sigma out}+P_{n\,out})]/[P_{s\,in}/(P_{i\Sigma\,in}+P_{n\,in})],$$

where P_s, $P_{i\Sigma}$, P_n -- power of signal, total interference power and thermal noise power, respectively, defined by the formulas:

$$P_{s\ out}=W^{+}_{opt}R_1W_{opt};\ P_{i\ out}=\sum_{j=2}^{J}W^{+}_{opt}R_jW_{opt};\ P_{n\ out}=W^{+}_{opt}R_nW_{opt}.$$

Here R_j, R_n -- correlation matrices of vectors of jth radiation and thermal noise, respectively; J -- number of emissions; W_{opt} -- optimal value of WCV in accordance with MMSD criterion, + -- Hermite conjugation sign. In this formula and the ones below radiation subscripted j=1 corresponds to the valid signal, and $j\neq 1$ corresponds to interference.

Assuming that the frequency spectra of the signal and interference coincide and that the fronts of the arriving waves are flat, the efficiency factor η for an equidistant antenna array is expressed through the signal and interference parameters and the array geometry as

$$\eta = \frac{P_{s\ out}\ W^{+}_{opt} S_1 W_{opt} \Xi(\alpha_1, \theta_1)}{\sum\limits_{j=2}^{J} \Xi(\alpha_j, \theta_j) P_j W^{+}_{opt} S_j W_{opt} + P_{n\ out} |W_{opt}|^2} \Big/ \frac{P_1}{\sum\limits_{j=2}^{J} P_j + P_{n\ in}}, \tag{7}$$

where P_j -- power of jth radiation; $\Theta(\alpha_j, \theta_j)$ -- directivity characteristic of array elements in direction of angles of arrival α_j, θ_j; S_j -- square nxn matrix with elements calculated by the formula

$$S_j(l_1, l_2) = \exp\left\{i \frac{2\pi d}{\lambda} \sin\theta_j [(k_1 - k_2)\cos\alpha_j + (m_1 - m_2)\sin\alpha_j]\right\},$$

where $\ell_1=(m_1-1)K+k_1$; $\ell_2=(m_2-1)K+k_2$; m_1, $m_2=\overline{1,M}$; k_1, $k_2=\overline{1,K}$; $\ell=\overline{1,L}$; $L=KM$; d -- distance between array elements; λ -- carrier wavelength.

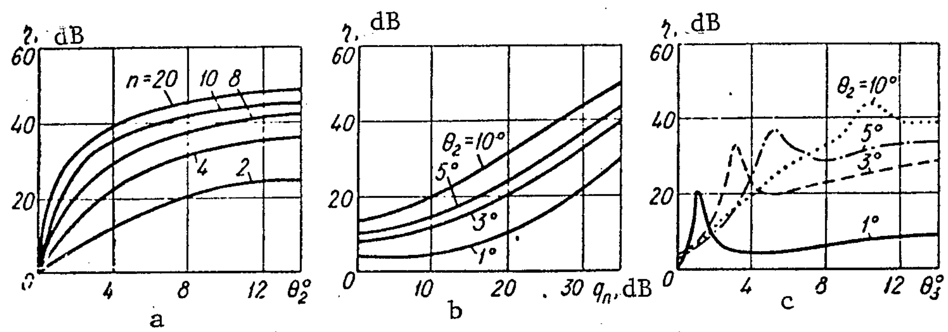


Fig. 2

If the signal/interference situation is stationary, the values of the WCV estimates entering into formula (7) can be found by solving Wiener-Hopf equation (1). It is entirely acceptable to use this formula for analyzing the effectiveness of space-time processing in the present case since the statement of the problem and the effectiveness criterion selected are the same for Kalman-Bussy and Wiener-Hopf procedures in this case, and it can be demonstrated that the effectiveness of the solutions obtained is the same for both procedures [5].

The effectiveness factor η was investigated in accordance with formula (7) as a function of the number of receiving elements n, the relative interference level $q_i=P_i/P_n$ and the angles of arrival of the interference. The results of the calculation were used to construct by computer the graphs shown in Fig. 2 a–c. The functions are calculated for $d=\lambda/2$ and isotropic array elements; The angle of arrival of the valid signal $\theta_1=0°$.

The following conclusions can be drawn from analyzing the findings. The value of the effectiveness factor η is a strong function of the angular mismatch θ_2 (Fig. 2a) when the interference is in the main lobe of the array pattern; the function is much weaker outside the main lobe. The relative attenuation of the interference increases as its power level; however, the absolute level of interference at the array output becomes greater. This can be seen from the plots in Fig. 2b, which are constructed for n=10 and different angles of arrival of interference θ_2. Figure 2c, for the case of two interference events, shows plots of η as a function of the angle of arrival of one interference event θ_3 with the other event θ_2, having fixed angles of arrival for n=10, with total power of both interference events $q_i=30$ dB, corresponding to the power of one noise event for the plots in Fig. 2a. Comparison of these functions indicates that the presence of additional interference reduces array functioning effectiveness significantly, especially when it falls in the main lobe of the pattern. When the angles of arrival of both interference events are the same, i.e., when $\theta_2=\theta_3$, the effectiveness factor increases to a level corresponding to the operation of one interference source with the same power as the combined power of the two sources. For some interference sources the behavior of the effectiveness factor has maxima when the interference events have the same angles of arrival. This is confirmed by the fact that when the number of interference events coming from different directions increases and the total power of all of the interference is equal, the effectiveness of space-time processing falls off.

In the general case, for p strong interference sources the dimensionality of the realization observed, i.e., the number of array elements, must be $n \geq p+1$ [6]. Furthermore, as analysis indicates, array effectiveness increases as n is increased when the number of interference sources is constant (Fig. 2a). The amount of interference attenuation does not increase uniformly. If there are few elements in the array, augmenting them makes it possible to increase the degree of attenuation significantly. When the array contains many elements, the gain from adding still more increases more slowly, while the time required for the processor to compute the optimal weight coefficients increases significantly.

The degree of interference attenuation in systems employing space-time signal processing can thus be very high, and depends upon the number of elements in the array, the number and power of interference sources and the angles of arrival.

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6900
CSO: 8144/1394

UDC: 517.951

DIFFRACTION OF ELECTRIC DIPOLE FIELD ON IDEALLY CONDUCTING SPHERE LOADED WITH RECTANGULAR SLOT

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian
Vol 26, No 3, Mar 83 (manuscript received 29 Jun 82) pp 344-350

GAVRIS, I. B., Belorussian State University

[Abstract] A study is made of the problem of modeling a radiating system consisting of an electric dipole source with arbitrary orientation of moment and an ideally conducting sphere loaded with a rectangular slot. The purpose of the work is to analyze the computer-generated results of a numerical solution of the problem, represented by a radiation pattern and graphs characterizing the behavior of the active and reactive components of conductivity of the load on the slot, which provides maximum radiation power for the system in question. By changing the orientation of the dipole moment, displacing a center of the slot along a meridian and a parallel, increasing the distance from the dipole to the surface of the sphere, and by changing the dimensions of the sphere and the slot it is possible to produce radiation patterns of various forms and directions including near omnidirectional patterns convenient for close range communications. Figures 4; references 5: 4 Russian, 1 Western.
[240-6508]

UDC: 537.87;621.371

THRESHOLD SPECIFICS IN DIFFRACTION OF ELECTROMAGNETIC WAVES ON PERIODIC HALF SPACE

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 30 Jun 82) pp 339-343

AYVAZYAN, Yu. M., SOZINOV, V. A., All-Union Scientific-Research Institute of Physical-Technical and Electronic Measurements

[Abstract] A study is made of the diffraction of an electromagnetic wave flowing from a free half space to strike a two-dimensionally periodic dielectric medium. It is assumed that the dielectric half space is two-dimensionally periodic in planes parallel to the division boundary of the media. The root nature of the specifics developing in the amplitudes and phases of the reflected waves in free space and the waves in the periodic medium upon appearance of waves in a new nature in either of the media can be demonstrated on the basis of such general concept as the law of conservation of energy, periodic and analytic properties of solutions, the existence of root threshold parameters and a system of orthonormalized natural waves in the two-dimensionally periodic dielectric structure. A solution derived in an earlier work for the case of a unidimensionally periodic medium is cited as an example clearly demonstrating the appearance of threshold root anomalies. References 7: 3 Russian, 4 Western.
[340-6508]

UDC: 621.371:551

MILLIMETER WAVE BEAM FLUCTUATION ON PATH WITH REVERSE REFLECTION

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 29 Dec 81) pp 271-275

ANDREYEV, G. A., ZAKHAROV, A. S. and KLEYN, A. G., Yaroslavl State University

[Abstract] An experimental study was made of the amplitude fluctuations of a bundle of millimeter waves over a path with back reflection under various weather conditions. The experiments were by a radar system. At the same time fluctuations in signal amplitude in the centimeter waveband were also studied. The wave path passed over slightly broken terrain. A corner reflector with an effective scattering area of 3700 m^2 at the 36.2 GHz band was located at a distance of 17 km on a hill 46 m in height. The radar station used was an MRL-1 pulsed weather radar station. The antenna was a parabolic reflector and a dual band horn radiator. Analysis of the results indicated that the maximum depth of chaotic modulation (GCM) reached 30%. The appearance of inverse layers at night resulted in deep fading in

the microwave band. The integral distribution of fluctuations of amplitude logarithm was near normal. Theoretical description of millimeter wave amplitude fluctuations in the smooth disturbance method approximation must consider fluctuations in the rate of transfer and the external scale of turbulence. The authors thank M. N. Kochegarov for conducting computer calculations and A. N. Latatuyeva for assistance in processing experimental data. Figures 5; references 21: 16 Russian, 5 Western (1 in translation).
[240-6508]

UDC 621.391

EXPERIMENTAL STUDY OF MULTIPLICATIVE MODEL DESCRIBING MULTIPLE IONOSPHERIC REFLECTIONS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 24 May 82) pp 74-76

MIRKOTAN, S. F., ZHURAVLEV, S. V. and KOSOVTSOV, Yu. N.

[Abstract] A method has been proposed for determining the signal-noise energy parameter of a signal partly scattered in the ionosphere on the basis of the multiplicative statistical model of multiple reflections. This method has been subsequently confirmed in an experiment with a single signal and vertical probing of the F2 layer (3-12 MHz, Moscow 8.00-22.00 local time). This model does not assume a priori that all successive reflections have the probability distributions. Measurements of the signal envelope and processing of the data for first and second reflections indicate that this model is adequate, whereas the Rice model is not adequate for a multistep ionosphere. The results also reveal a shift of the probability density distributions toward lower values, possibly a consequence of scattering of the wave by the "rough" earth surface. Figures 4; references 6: 5 Russian, 1 Western (in translation).
[226-2415]

UDC 621.396.67

EFFECT OF FLUCTUATIONS OF PARAMETERS CHARACTERIZING MOMENTARY INTERFERENCE ON EFFICIENCY OF SPACE-POLARIZATION PROCESSING OF SIGNALS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 12 Aug 82) pp 71-74

RODIMOV, A. P., NIKITIN, S. V. and NIKITCHENKO, V. V.

[Abstract] Space-polarization processing of signals with momentary interference is analyzed, specifically the generating of weight factors on the

basis of preceding interference pulses. Inevitable fluctuations of the interference parameters from pulse to pulse, particularly of the direction from which the interference arrives, cause the vector of weight factors to deviate from the optimum and consequently to reduce the processing efficiency. The effect of such fluctuations is evaluated here in terms of loss in signal-to-(interference+noise) ratio. Assuming that the vector of weight factors is generated on the basis of the mean values of interference space-polarization parameters, the results indicate that the best way to minimize this effect is to detect the interference in the line of antenna elements and to orientate the antenna array accordingly. Figures 4; references 7: 6 Russian, 1 Western (in translation).
[226-2415]

UDC 621.396.677

ELECTRODYNAMIC ANALYSIS OF REFLECTOR AS ARRAY OF PARALLEL CONDUCTORS (CASE OF E-MODE WAVES)

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 29 Mar 82) pp 12-16

INSPEKTOROV, E. M.

[Abstract] An antenna reflector is regarded as an array of parallel cylindrical conductors and its near field is calculated, approximately, by replacing such an array with an equivalent surface whose impedance depends on the array parameters and the Vaynshteyn-Sivov boundary conditions. The surface current is determined from the corresponding Fredholm integral equation of the second kind, taking into account diffraction at the reflector edges. The problem is solved as a two-dimensional one for waves polarized in the E-plane, using a Cartesian system of coordinates and a curvilinear orthogonal one with a common z-axis. The contour is subdivided into N segments so that the integral equation reduces to a system of algebraic ones, the latter solvable by numerical methods. The integral of the kernel can be evaluated accurately for the nondiagonal coefficients ($p \neq q$) but only approximately according to a heuristic relation for the diagonal coefficients ($p \to q$), because it is divergent here (p- point where current is calculated, q- point where numerical integration is performed). The procedure has been applied to two models of a reflector, a plane array with the dimension 3.9λ and with edges rounded to a diameter of 0.016λ and a plane array with the dimension 4.4λ and with edges rounded to a diameter of 0.25λ. The results of calculations for the near field in the shadow region reveal a compensating effect, the field suppression in this region becoming more effective with an increasing intensity of the diffracted field and with an increasing transmission coefficient. Figures 4; references: 4 Russian.
[225-2415]

UDC 621.396.68

COMPRESSION OF PULSES STRETCHED BY ANTENNA ARRAY WITH OPTICAL EXCITATION

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 27 Dec 82) pp 77-79

MAKSAKOV, N. F.

[Abstract] A feedthrough antenna array with optical excitation and without scanning is considered. The unit consists of a primary element and a reradiating structure of transceiver elements with interconnection of them by uncontrolled feedthrough phase shifters. These phase shifters transform a plane wavefront into a spherical one for reception and a spherical wavefront to a plane one for transmission. An analysis of receiver operation reveals a stretching of short pulses. A method of compensating this is proposed which involves space ooding of the oscillation phases in a coding matrix and subsequent time compression of the phase-code-keyed signal in a matched filter. The latter constitute a delay line connected to another array of phase shifters, those with a code which is a mirror image of the phase-code-keyed signal. Figures 2; references: 5 Russian.
[226-2415]

UDC: 628.947:771.44

PARABOLIC-CYLINDRICAL REFLECTORS WITH LONGITUDINAL BENT CHANNELS FOR MOTION PICTURE LIGHT SOURCES

Moscow TEKHNIKA KINO I TELEVIDENIYA in Russian No 5, May 83 pp 38-41

SEMENIKHIN, N. T., All-Union Scientific Research Institute of Motion Picture Photography

[Abstract] Parabolic-cylindrical reflectors are used in modern motion picture light sources. One shortcoming of these reflectors is the variation in lighting characteristics of instruments of the same type caused by the instability of the process of shot peening used to work the surfaces of the reflectors. More stable light parameters can be achieved by creation of optical elements in the form of cylindrical channels of various widths and radiuses in order to produce the necessary additional light scattering in the vertical plane. Diagrams illustrate the paths of rays created by such light reflectors. The method of designing the geometric parameters of optical elements of reflectors embodied in the equations and diagrams presented in this article allow calculations to be performed with an accuracy of 10 to 15%. Reflectors having channels with identical parameters achieve practically complete reproducibility of the desired lighting characteristics, with variations of 2 to 5%. Figures 5; references: 3 Russian.
[246-6508]

UDC: 778.5(47+57)"1982"

REVIEW OF MAJOR WORKS ON PROFESSIONAL CINEMATOGRAPHIC TECHNIQUES OF 1982

Moscow TEKHNIKA KINO I TELEVIDENIYA in Russian No 5, May 83 pp 3-20

YEMEL'YANOV, G. F., OVSYANNIKOVA, N. A., ARNOLD. R. R., BOLOTNIKOV, I. M., DOYNIKOV, B. N., GOLOSTENOV, G. A., KRASNIKOV, S. F., ALMAZOV, V. Ye., BERKENGEYM, A. B., VELICHKO, G. V., Central Scientific and Technical Information Office, National Institute for Cinematography, and BONDARCHUK, V. M., IRZ, P. V., GILINSKIY, A. G. and DROZDOV, V. M., "Ekran" Scientific and Production Organization

[Abstract] Major developments in Soviet cinematographic hardware for 1982 are described. Photographs are presented of most of the items of equipment; technical characteristics are given for a few. Hardware discussed includes the KPU standardized power supply system, 2GSP gyrostabilizer, FPR-3 projector, a 28 millimeter lens, the "Saturn-4000-35" spotlight, the PKS-1000 underwater cinematographic light source, K90K-41 sound control and mixing panel with 32 microphone inputs and 18 recording output channels, the KZM-28 recording system, various multitrack magnetic head units, earphones, taperecorder drive control units, the A742V field sound track application stage, an automated print maker and light sources for print makers, the DKsSh-4000 xenon projection lamp, the DPT-2.5 theater slide projector, splicing blocks for 16 and 70 mm film and the box-S housing for underwater photography. Figures 21; tables 2.
[246-6508]

UDC: 778.533.1

CLASSIFICATION OF FILM TRANSPORT MECHANISMS

Moscow TEKHNIKA KINO I TELEVIDENIYA in Russian No 5, May 83 pp 21-26

LEVITIN, G. V.

[Abstract] A classification of film transport mechanism hardware is presented, based primarily on the nature of the connection between the individual units in the film transport mechanism. The individual units are connected by two methods: by the film itself and by the elements of the drive mechanism which transmit rotation among the units. Film transport mechanisms designed for use in motion picture cameras and projectors as well as developing machines are considered. An extensive hierarchical block diagram of film transport mechanisms is presented, covering 8 mm through 70 mm formats. The task which remains is one of optimal synthesis of related film transport mechanisms based on an overall theory yet to be developed considering the interrelationships and mutual influence of individual film transport mechanism components. Figures 12; references: 13 Russian.
[246-6508]

UDC: 778.534.4:78

MODERN MOTION PICTURE STUDIO MUSIC RECORDING SYSTEM

Moscow TEKHNIKA KINO I TELEVIDENIYA in Russian No 5, May 83 pp 33-38

TSUKERMAN, M. Ya., "Mosfil'm" Motion Picture Studio

[Abstract] In 1981 a modern motion picture sound recording system employing many complex and very expensive elements embodying all the latest achievements in electronics was installed at the "Mosfil'm" motion picture studio. This system is used as an example in order to discuss briefly modern methods of recording music for the motion picture industry. The system consists of four main parts: the studio area with microphones, mixer panel, magnetic tape recorder system and overall control system. The studio has a volume of 10,000 cubic meters, reverberation time about 1.8 s. As many as 28 microphones can be connected to the mixer panel at once, though other mixers have as many as 56 inputs, because ideally a separate microphone is used for each instrument or small group of instruments. At times several microphones are used for one instrument, as in the case of percussion instruments. Typical recording quality figures include a noise level at the input of -128 dB, at output 18 dB, frequency range 20 to 20,000 cycles, zero phase distortion. Automatic equipment records the manipulations of the many controls of the mixer panels by the sound operator during the first pass of recording of a piece of music, then reproduces them precisely, leaving it to the operator merely to make find adjustments to the settings desired as the piece is played. Use of such modern equipment has more than doubled the productivity of labor in recording of motion picture sound tracks. Figures 4.
[246-6508]

UDC 621.316.729

WIDEBAND SYNCHRONIZER FOR MICROWAVE OSCILLATORS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 17 Sep 82) pp 36-37

KONSTANTINOV, V. I., MASALOV, V. L., TOKAREV, A. D. and FILYUK, A. A.

[Abstract] A synchronizer for microwave oscillators with phase-locking automatic frequency control has been developed which consists of a 1-12 GHz Ch3-38 frequency meter with transducer and quartz reference oscillator, a wideband mixer, an external heterodyne, also an i-f amplifier, two phase detectors, a buffer amplifier, two d.c. amplifiers with low-pass filters, and a generator of linearly varying voltage. The frequency instability of the quartz oscillator is within 10^{-10} for short periods (1 s) and within 10^{-9} for long periods (1 h). The frequency meter mixes the signal from the working oscillator at frequency f_0 with a harmonic from the quartz oscillator at frequency f_{ref}. Its output signal at the difference frequency $f_1 = f_0 - f_{ref}$ is amplitifed in the transducer and through an emitter follower proceeds to one input of the wideband mixer, at the other input of which there appears a signal from the external heterodyne at a frequency $f_2 = f_1 \pm 5$ MHz. The frequency of the working oscillator is maintained at $f_0 = f_{ref} \pm f_2 \pm 5$ MHz during synchronism and can be varied, while being continuously measured by the frequency meter, through tuning of the external heterodyne by means of the generator of linearly varying voltage with amplitude and duration control. Figures 2; references: 3 Russian.
[226-2415]

UDC 621.317.729.2

MEASUREMENT OF HIGH-FREQUENCY PARAMETERS OF SLOWDOWN CIRCUITS IN O-TYPE MICROWAVE ELECTRON DEVICES

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 25 Jan 82) pp 89-91

RACHKOV, V. A.

[Abstract] The method of small attenuations with an electron-sheath probe is proposed for measurement of the coupling resistance and the retardation coefficient in the slowdown circuit of completely assembled O-type microwave electron devices such as traveling-wave or backward-wave tubes with a local absorber or a gap in that slowdown structure. The gist of this method is applying a microwave signal to the output of the device and measuring, through a directional coupler, the electronic gain when interaction of the electron beam and the wave reflected by the absorber or the gap in the slowdown structure occurs. The method was tried successfully on various commercial traveling-wave tubes in attest stand containing, in addition to a directional coupler, three ferrite diodes, two detectors, a polarizational attenuator, and an amplifier with oscilloscope as an interaction indicator. The readings can be processed on a small computer such as the MIR-2 or with the aid of graphs and tables. Figures 3; references 5: 4 Russian, 1 Western. [225-2415]

UDC 621.372

ALGORITHMS OF PROCEDURE FOR IMPROVING ACCURACY OF SELECTIVE SIGNAL CONVERTERS

Kiev ELEKTRONNOYE MODELIROVANIYE in Russian Vol 5, No 3, May-Jun 83 (manuscript received 3 Nov 82) pp 48-51

MASLAKOV, GENRIKH NIKOLAYEVICH, candidate of technical sciences, head of Zhitomir Affiliate, Special Design and Technological Office, Institute of Problems of Power Engineering Modeling, UkrSSR Academy of Sciences

[Abstract] The accuracy of selective signal converters under conditions of a priori indeterminacy, with additive and multiplicative processing errors, can only be improved by an adaptive procedure. Three algorithms of such a procedure are proposed for improving the converter accuracy in this case. The first algorithm is for selection of the optimum converter transfer function for a given signal, depending on the signal spectrum distribution over a given frequency range. Its main advantage is the direct use of information about the current signal spectrum and immediate decision about the transfer function, following a parallel analysis of all conjunctions of indicators. Its disadvantage is a stiff deterministic equation making the algorithm inefficient when the frequency range has been subdivided into only a few wide subranges. The 3-step second algorithm either makes the optimum selection of transfer function initially on the basis of an experimental

analysis of indicators, without intermediate selections, or determines a better classification at the first level. The algorithm eliminates full scan of all classifications at first and second levels, through parallel analysis of all indicator combinations on the first step. The third algorithm is a transformation of the first one, from which it differs in that the selection of the optimum transfer function is decided when at least one of the noninformative indicator equals unity. References 7: 6 Russian, 1 Western (in translation).
[238-2415]

UDC 621.372.6

PARAMETRIC SENSITIVITY OF MICROWAVE CIRCUITS

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received, after completion, 8 Jun 82) pp 41-45

GENIS, Ye. A., ZAIKIN, B. M. and KAZANDZHAN, N. N.

[Abstract] The sensitivity of linear lumped-parameter microwave cirauits to variation of their parameters is calculated, for design purposes, with the aid of generalized scattering parameters in the inverse admittance matrix. All algorithms are based on solution of the system of equations $WX = C$ (X- vector containing the normalized amplitudies of incident and reflected waves in inner and outer branches, W- matrix of the scattering coefficients of multipole circuit components with information about circuit topology, C- vector of amplitudes of the waves from driving oscillators). The algorithms for calculating first-order and second-order sensitivity are constructed on this basis, with the corresponding partial derivatives converted to single and double sums respectively. They are applied, for illustration, to a cascade network of three six-pole rings with $X = [\alpha_1 b_2 b_3 b_4 b_5 b_6 \alpha_7 \alpha_8 b_9 b_{10} \alpha_{11} \alpha_{12} b_{13} b_{14}]^T$. The derivatives $\partial S_{137}/\partial r_B$ (r_B- lumped resistance in one of the two outer rings) is calculated as an example. Figures 2; references 9: 6 Russian, 3 Western.
[225-2415]

UDC 621.372.54

SYNTHESIS OF NONRECURSIVE FILTER WITH RHOMBIC PASS REGION

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 9 Feb 82) pp 21-24

BORODYANSKIY, A. A.

[Abstract] A two-dimensional integrating filter is synthesized for recovery of two-dimensional messages from readings taken at the corners of a triangular array. Such a message, with a bounded spectrum is discretized and processed as a numerical data array. The characteristic of an ideal discrete filter for this purpose is represented in the form of a double trigonometric series with a weighting window which improves the convergence, Fourier se ies being unsuitable here because of the Gibbs effect. A comparison of Hemming and Feuer window indicates that a latter is more efficient in terms of the number of readings and more simple in terms of practical realization for message processing in real time at less than high speed. Figures 4; references 5: 2 Russian, 3 Western (2 in translation).
[225-2415]

UDC 621.372.83

WIDEBAND POWER DIVIDER-ADDERS WITH CHANNEL REDUNDANCY

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 10 Jun 81) pp 68-70

GORBACHEV, A. P. and NEVEROV, S. G.

[Abstract] Microwave power divider-adders with channel redundancy have been developed using quarter-wavelength directional couplers. The can add the power of $N = 2^m$ ($m = 1,2,\ldots$) identical amplifiers A_i ($i = 1,2,3,\ldots,N$) in normal operation with simultaneous standby for each in three different modes of failure. The switches are controlled automatically by a device which generates output signals depending on the absence or the presence and magnitude of a difference voltage across ballast resistors. The bandwidth of such a power divider-adder is evaluated with the aid of a 2Nx2N square partitioned scattering matrix. On the basis of this performance analysis a method is devised for design optimization with respect to frequency characteristics. A 4-channel divider-adder was built with coupled symmetric strip lines, shielded on the wider sides, for adding the power of 3 W amplifiers over a frequency range with a 1.5 overlap factor. The design was optimized for a 50 ohm total channel impedance and the microwave part was produced by the photochemical process on a 0.15 mm thick sheet of FAF-4 material between two standard 1.5 mm thick plates of the same material. Figures 3; references 7: 6 Russian, 1 Western.
[225-2415]

UDC 621.373.144.1:681.335.7

POWER PULSE INTEGRATOR

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 4 Dec 82) pp 90-91

BONDAR', V. A., TOPOR, A. V. and KAS'YANOV, A. U.

[Abstract] A device has been designed for integrating power pulses of intricate waveform. Its principal components are two operational amplifiers, one followed by a high-voltage output power amplifier and the other in the common feedback loop, a shaping capacitor, a charging transistor and a discharging transistor, as well as a matching transistor amplifier. The main operational amplifier minimizes nonlinear distortion of the output signal and the negative current feedback minimizes errors of integration. Altogether there are nine transistors, including those in the fontrol module across the input of the pulse shaping stage. The device can shape and integrate saw-tooth pulses with linear, quadratic, cubic, or other voltage rise as well as triangular, trapezoidal, and more intricate pulses. It can be used in modulators, converters, measuring instruments, and for radio signal or video signal processing. Figures 2.
[226-2415]

UDC 621.373.423

OPTIMIZATION OF OPERATING MODE OF AUTODYNES WITH GUNN-EFFECT DIODES

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 28 Sep 82) pp 30-33

TERESHCHENKO, A. F. and DEKIN, G. N.

[Abstract] The operation of autodynes with Gunn-effect diodes is analyzed, for the purpose of their performance optimization. The current-voltage characteristic is described by the approximating equation $y = A(x'e^{1-x'} + B)$, $y = (i_d - I_{min})/\Delta I$, $x' = 1 + \gamma(x - 1)$, $x = (v - V_{min})/\Delta V$, $\Delta I = I_{max} - I_{min}$,

$\Delta V = V_{crit} - V_{min}$, $A = [1 - (1 - \gamma)e^{\gamma}]^{-1}$, $B = (\gamma - 1)e^{\gamma}$, i_d- diode operating voltage, I_{min} and V_{min} minimum current and minimum voltage on descending branch of current-voltage curve, V_{cr}- critical voltage at which oscillation starts, γ- quality parameter based on comparison of theoretical and experimental efficiency. The diode performance in autonomous mode and in the autodyning mode is calculated on this basis. The results reveal a dependence of the voltage transfer ratio, which characterizes the sensitivity, on the operating point and on the magnitude of the high-frequency load. For optimum autodyne performance it is necessary to ensure first a sufficiently large stability margin in the autonomous mode so as to avoid exciting low-frequency

oscillations, and then operate below that stability limit at the maximum transfer ratio with a load resistance equal to the dynamic resistance of the diode in the autonomous mode. This is illustrated on an antenna with a high-frequency load and a low-frequency load. Figures 2; references: 9 Russian.
[226-2415]

UDC 621.373.5

SOME PROBLEMS IN DESIGN OF MINIATURE TUNNEL-DIODE OSCILLATORS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 28 Sep 82) pp 27-30

REZNEV, A. A.

[Abstract] Problems in the design of miniature low-power tunnel-diode oscillators at the hybrid integration level arise in the realization of high-Q tank circuits with adequate stability at the fundamental frequency. The Q-factor under load of a high-frequency oscillator is calculated here on the basis of a simple two-loop equivalent circuit with only three passive elements and a fifth-degree polynomial approximation of the current-voltage characteristic, with the amplitude of steady-state oscillations found from the solution to the corresponding nonlinear equation. The Q-factor under load of a microwave oscillator is determined similarly on the basis of a more intricate five-loop equivalent circuit with seven passive elements. The temperature stability of the oscillator frequency is estimated in each case, assuming that only the diode capacitance is sensitive to destabilizing temperature changes. Numerical data characterizing the dependence of the frequency drift on the temperature drift and of the Q-factor on the bias voltage are shown for 1I401A and 1I104D tunnel diodes. Figures 2; tables 1; references 7: 5 Russian, 2 Western (both in translation).
[226-2415]

UDC 621.384

REGULATION CHARACTERISTICS OF PULSE CONVERTER FOR HIGH-VOLTAGE SOURCE

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 13 Jul 82) pp 92-94

PAVLOV, S. V., SAKHAROV, V. A. and VASIL'YEVA, N. K.

[Abstract] A pulse converter for a high-voltage source is described which includes a TVS-90LTs step-up transformer (1:30-1:40) with an air gap and a voltage rectifier-multiplier across the secondary. The 200 μH leakage

inductance of the secondary winding ensures an oscillatory discharge of the 0.025 μF capacitor across the thyristor-controlled (250-400 Hz) primary winding and in this way a stable thyristor cutoff during half the period. The voltage regulation characteristics, including the maximum regulation factor, as well as the efficiency and other performance parameters are calculated on the basis of a simple equivalent circuit with capacitance-coupled primary and secondary and with an ideal switch across the secondary. Figures 5; references: 2 Russian.
[226-2415]

UDC 621.391.822:621.375

NOISE FACTOR OF DEVICES WITH FEEDBACK

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 11 Sep 82) pp 22-24

KHEVROLIN, A. V.

[Abstract] A device is considered consisting of a noisy amplifier with a directional coupler in series, the latter returning a fraction of the output power to the input through a noisy feedback stage with a delay line. All components are matched with respect to input and output impedances, are independent, and operate independently in the linear-signal mode. The overall gain of the device and the noise factor at its output are calculated for the two extreme cases of noncoherent and coherent superposition at the amplifier input. In the first case, generally with a wideband incoming signal, the signal-and-noise correlation time is much shorter than the feedback delay time. In the second case, generally with a narrow-band incoming signal, the signal-and-noise correlation time can be much longer than the feedback delay time. The results of calculations, performed by the method of equivalent generators, reveals that in each case a feedback increases the overall noise factor by an amount which does not depend on the mode of feedback. The overall noise factor can be decreased by decreasing the gain and the noise factor of the feedback circuit. Figures 4; references 3: 2 Russian, 1 Japanese.
[226-2415]

UDC 621.391.2

PERFORMANCE ESTIMATION FOR AM SIGNAL FILTERING AGAINST BACKGROUND OF WHITE NOISE AND AM INTERFERENCE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 4, Apr 83 (manuscript received 5 Jan 82) pp 96-98

[Article by P.N. Serdyukov and V.P. Burdzeyko]

[Text] Aspects of optimal reception of AM signals against the background of white noise and narrowband interference were examined in [1,2], where the structural diagrams of optimal receivers were synthesized. However, [1] provides no analysis of filtering performance, and [2] concludes that narrowband interference has no influence on reception noise tolerance.

On the basis of Markov's theory of nonlinear filtering, the present study analyzes AM signal reception performance against the background of white noise and AM interference for an a priori assigned Markov model of the parameters of signal and interference over a wide range of variation in the parameters.

We write the signal input to the receiver in the form

$$y = S + V + n, \tag{1}$$

where $S=S(t)=A(1+m_1\lambda)\sin(\omega_1 t+\phi)$ -- valid signal; $V=V(t)=B(1+m_2\mu)\sin(\omega_2 t+\Phi)$ -- interference; $n=n(t)$ -- white noise with characteristics $\overline{n(t)}=0$; $n(t)n(t+\tau)=$ $=0.5N\delta(\tau)$, λ, μ -- modulating functions of valid signal and interference with corresponding modulation indexes m_1 and m_2; Z, B, N, ω_1, ω_2 -- certain constants.

The remaining parameters which enter into (1) are assigned by these a priori equations, which are characteristic for describing messages in radio communications:

$$\left.\begin{aligned} d\lambda/dt &= -h\lambda - h_1\lambda + \sqrt{b_\lambda}\xi_1 \\ d\lambda_1/dt &= -h_1\lambda_1 + \sqrt{b_\lambda}\xi_1 \\ d\varphi/dt &= \sqrt{B_\varphi}\xi_2 \end{aligned}\right\}, \tag{2}$$

$$\left.\begin{aligned} d\mu/dt &= -h_\mu\mu + \sqrt{b_\mu}\xi_3 \\ d\Phi/dt &= \sqrt{B_\Phi}\xi_4 \end{aligned}\right\}. \tag{3}$$

Here h, h_1, h_μ, b_λ, b_μ, B_ϕ, B_Φ -- certain constants; $h+h_1$ -- bandwidth of energy spectrum of process (λ, λ_1) at 0.5 level; h_μ -- bandwidth of energy spectrum of process μ; b_λ, b_μ, B_ϕ, B_Φ -- coefficients of diffusion of processes (2), (3); $\xi_i=\xi_i(t)$ -- white noise-type processes;

$$\bar{\xi}_i = 0, \qquad \overline{\xi_i\xi_{i\tau}} = \delta(\tau), \qquad \overline{\xi_i\xi_{j\tau}} = 0, \qquad i \neq j, \qquad i, j = \overline{1, 4}.$$

The quality with which the estimated parameters are filtered is determined by solving this matrix differential equation [3]:

$$\frac{d}{dt}K = \left[\begin{array}{c|c} H & 0 \\ \hline 0 & U \end{array}\right] K + K \left[\begin{array}{c|c} H & 0 \\ \hline 0 & U \end{array}\right]^{T} + \left[\begin{array}{c|c} G & 0 \\ \hline 0 & Q \end{array}\right] - K\Phi K, \tag{4}$$

where $\left[\begin{array}{c|c} H & 0 \\ \hline 0 & U \end{array}\right]$, $\left[\begin{array}{c|c} G & 0 \\ \hline 0 & Q \end{array}\right]$ -- block matrices; T -- transposition sign;

K -- normalized matrix of a posteriori cumulants; $\Phi = \left\| \frac{\partial^2 F}{\partial x_i \partial x_j} \right\|$ --

square matrix, where the function F for observation (1) appears as [3] $F=(1/N)[2y(\hat{S}+\hat{V})-(\hat{S}+\hat{V})^2]$, where $\hat{S}$, $\hat{V}$ -- estimates of signal and interference; x_i, x_j -- components of estimated parameter vector.

Dividing (4) by h from the left and right, we write the following initial matrices for the condition $h=h_1$:

$$H=\begin{Vmatrix} -1 & -\sqrt{2} & 0 \\ 0 & -1 & 0 \\ 0 & 0 & 0 \end{Vmatrix}; \quad U=\begin{Vmatrix} -h_\mu/h & 0 \\ & \\ 0 & 0 \end{Vmatrix};$$
$$G=\begin{Vmatrix} 4 & 2\sqrt{2} & 0 \\ 2\sqrt{2} & 2 & 0 \\ 0 & 0 & D_\varphi \end{Vmatrix}; \quad Q=\begin{Vmatrix} 2h_\mu/h & 0 \\ & \\ 0 & D_\Phi \end{Vmatrix}; \tag{5}$$

D_Φ, D_ϕ -- increment in phase of oscillators during signal correlation time; $D_i=B_i/h$, $i=\phi,\Phi$.

The matrix Φ appears as

$$\begin{Vmatrix} \Phi_{11} & 0 & 0 & \Phi_{14} & \Phi_{15} \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & \Phi_{33} & \Phi_{34} & \Phi_{35} \\ \Phi_{41} & 0 & \Phi_{43} & \Phi_{44} & 0 \\ \Phi_{51} & 0 & \Phi_{53} & 0 & \Phi_{55} \end{Vmatrix}, \tag{6}$$

where

$$\Phi_{11}=R_s m_1^2; \quad \Phi_{14}=\sqrt{R_s R_i}\, m_1 m_2 \cos(\omega_p t+\hat{\theta});$$
$$\Phi_{33}=R_s(1+m_1^2); \quad \Phi_{34}=-\sqrt{R_s R_i}\,(1+m_1\hat{\lambda})\, m_2 \sin(\omega_p t+\hat{\theta});$$
$$\Phi_{44}=R_i m_2^2; \quad \Phi_{15}=\sqrt{R_s R_i}\,(1+m_2\hat{\mu})\, m_1 \sin(\omega_p t+\hat{\theta});$$
$$\Phi_{55}=R_i(1+m_1^2); \quad \Phi_{35}=\sqrt{R_s R_i}\,(1+m_1\hat{\lambda})(1+m_2\hat{\mu}) \cos(\omega_p t+\hat{\theta});$$

$R_s = A^2/Nh$ -- signal/noise ratio; $R_i = B^2/Nh$ -- interference/noise ratio in signal band.

When the a posteriori accuracy is high, the estimates are close to the actual parameter values.

Then, after averaging, we obtain

$$\begin{aligned} \tilde{\Phi}_{15} &= \sqrt{R_s R_i}\, m_1 \sin(\omega_p t + \hat{\Theta}); \\ \tilde{\Phi}_{34} &= \sqrt{R_s R_i}\, m_2 \sin(\omega_p t + \hat{\Theta}); \\ \tilde{\Phi}_{35} &= \sqrt{R_s R_i} \cos(\omega_p t + \hat{\Theta}), \end{aligned} \tag{7}$$

where ω_p -- difference between carrier frequencies of signal and interference; $\hat{\Theta} = \hat{\phi} - \hat{\Phi}$ -- estimate of random initial phase.

Substituting (5) and (6), considering (7), in (4), we solve this equation numerically in normalized time $t^* = ht$, when the initial conditions $K_{ij} = 1$ for $i = j$ and $K_{ij} = 0$ for $i \neq j$. The normalized difference $\omega_n = \omega_p/h$ and the noise and signal band ratio $h_n = h_\mu/h$ are used as additional parameters.

By averaging the solutions of equation (4), we obtain the sought values of the normalized accumulators. The values of the random initial phase are formed by a uniform number generator on the interval $(0-2\pi)$.

The results of the solutions shown for relative a posteriori distribution $\sigma_\lambda^2 = K_{\lambda\lambda}/\delta^2_{\lambda pr}$ of the filtered parameter λ indicate that the influence of random initial phase on the process of establishing the a posteriori dispersion is insignificant (Fig. 1). Here $\delta^2_{\lambda pr}$ -- a priori dispersion of λ. Results are then cited which are averaged over three realizations (L=3).

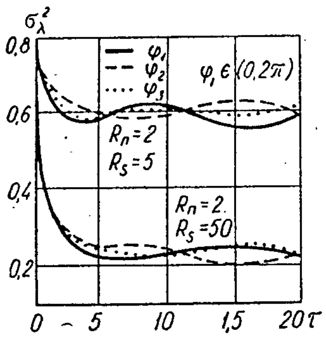

Fig. 1

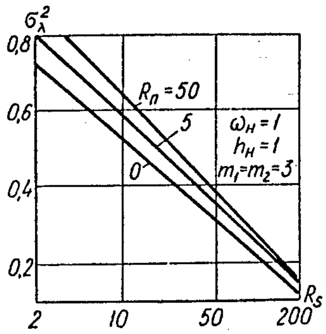

Fig. 2

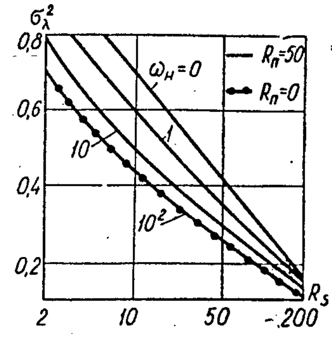

Fig. 3

Filtering performance drops off as interference increases (Fig. 2). When the interference level is constant the increase in the distance between the carrier frequencies of the signal and interference reduces the filtering error, which when ω_n is substantial corresponds to reception performance against the background of white noise (Fig. 3).

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ANALYSIS OF ACCURACY AND SENSITIVITY OF ALGORITHMS FOR PROCESSING MULTI-DIMENSIONAL SIGNALS AND INTERFERENCE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 4, Apr 83 (manuscript received 21 Jun 82) pp 93-96

[Article by V.V. Popovskiy, Ye.I. Glushankov and B.V. Voronkov]

[Text] In radio communications, radar, radiometry and radio navigations it is often necessary to use algorithms for processing multidimensional signals and interference, such as algorithms for combined estimation of different parameters and space-time processing algorithms [1,2], among others. The use of the variable-state method [3] in these cases makes it possible to obtain optimal linear and nonlinear algorithms which can be realized and which operate in real time. As the dimensionality of the observation space (N) increases, the complexity of the algorithms increases sharply, so that the number of differential equations (ℓ) which must be solved becomes equal to $\ell=N(N+3)/4$ [3]. At the same time, it is sometimes possible to simplify the algorithm by discarding the mutual cumulant, which for N=40 makes it possible to throw out 340 of 430 equations right away.

In many cases there are N measurements of the same process x(t); therefore, it is interesting to estimate how processing performance improves as n increases. It is also interesting to evaluate the sensitivity of multi-dimensional algorithms, since the actual signal-interference situation in practice does not always correspond to the parameters of the selected statistical model of the signals and interference.

Let the multidimensional models of the state x(t) and observation z(t) of the valid signal be assigned in the form of the linear difference equations

$$x(k)=F(k)\,x(k-1)+G(k)\,w(k),\quad z(k)=H(k)\,x(k)+v(k), \qquad (1)-(2)$$

where x(k) -- n-dimensional state vector; z(k) -- m-dimensional observation vector; F(k), G(k), H(k) -- certain matrices; dimensionality of F(k) and G(k) -- nxn; dimensionality of H(k) -- mxn; w(k), v(k) -- white noise with null mean

and known covariation matrices: cov $\{w(k), w(i)\}=V_w(k)\delta(k-i)$, cov$\{v(k), v(i)\}=V_v(k)\delta(k-i)$, cov$\{w(k),v(i)\}=0$.

The recursive algorithm for extimating x(k) by observations z(k) is assigned by these relationships [3]:

$$\hat{x}(k) = F(k)\hat{x}(k-1) + K(k)[z(k) - H(k)F(x)\hat{x}(k-1)], \tag{3}$$

$$K(k) = V_{\tilde{x}}(k, k-1)H^T(k)[H(k)V_{\tilde{x}}(k, k-1)H^T(k) + V_v(k)]^{-1} \tag{4}$$

$$V_{\tilde{x}}(k, k-1) = F(k)V_{\tilde{x}}(k-1)F^T(k) + G(k)V_w(k)G^T(k), \tag{5}$$

$$V_{\tilde{x}}(k) = [I - K(k)H(k)]V_{\tilde{x}}(k, k-1), \tag{6}$$

$$\hat{x}(0) = \langle x(0)\rangle,\ V_{\tilde{x}}(0) = V_{\tilde{x}}(0, 0) = V_x(0), \tag{7}$$

where $\hat{x}(k)$ -- estimate of x(k), $\tilde{x}(k)=x(k)-\hat{x}(k)$ -- estimation error; K(k) -- weight vector; $V_{\tilde{x}}(k)$ -- a posteriori error dispersion;

I -- identity matrix; r -- transposition sign; <•> -- statistical averaging. The accuracy with which the vector x(k) is estimated by observations z(k) and the sensitivity of the algorithms are analyzed by the value of the a posteriori estimate error dispersion $V_{\tilde{x}}(k)$.

We shall analyze the accuracy of the algorithms for processing multi-dimensional signals and interference with a signal power to noise power ratio in the receiving band $P_s/P_n=10^2$. Figure 1a shows the results of the analysis. The solid lines indicate the relationships for algorithms which allow for cross-correlation (cross-cumulants). The dotted line represents the relationship for a four-dimensional (n=m=4) algorithm which does not allow for cross-cumulants. The results indicate that the gain in accuracy of processing algorithms which allow for cross-cumulants increases as the dimensionality increases. Therefore, when the dimensionality of the system is substantial (n=m), algorithms which allow for cross-cumulants provide better accuracy in estimating the process x(t). At the same time, increasing n and m has no effect on the accuracy of algorithms which do not allow for cross-cumulants.

Figures 1b and c show the results of analyzing the sensitivity of processing algorithms to deviation in the parameters of the selected model from the actual signal-interference situation. We shall not be using formula (6) to calculate the a posteriori estimation error dispersion, since it does

not allow for instability and possible divergence of the algorithm when the observation noise level v(k) is sufficiently small [3], as well as on the basis of the real estimation errors $\tilde{x}(k)=x(k)-\hat{x}(k)$ obtained for 40 identical statistically independent models. The dispersion $V_{\tilde{x}}(k)$ is determined in each step k by the formula [4]*:

$$V_{\tilde{x}}(k) = \frac{1}{M-1} \sum_{i=1}^{M} [x_i(k) - \hat{x}_i(k)]\,[x_i(k) - \hat{x}_i(k)]^{\tau}. \tag{8}$$

The spectral density of the observation noise power V_v and correlation coefficient τ_{cor} of the real process is most often observed to deviate in practice from the parameters of the model. The solid lines in the figures show the functions for deviations of V_v, and the dotted lines show the deviations for τ_{cor}**.

The results of analyzing the sensitivity of the two-dimensional algorithm, disregarding cross-cumulants, are shown in Fig. 1b for different signal/noise ratios. It is apparent from the figure that for high signal/noise ratios ($P_s/P_n=10^4$) the algorithm is more sensitive to deviation of the model parameters from the actual situation than for $P_s/P_n=10^2$. This is explained by the fact that with high P_s/P_n the instability of the algorithm increases because of the influence of small observation noise.

Figure 1c illustrates the results of analyzing the sensitivity of two-dimensional algorithms which do (curves 3,4) and do not (curves 1,2) allow for cross-cumulants with signal/noise ratio $P_s/P_n=10^2$. It is apparent from the figure that the algorithm which allows for cross-cumulants is less sensitive to deviation in the parameters of the selected statistical model from the actual signal-interference situation. The relative estimation error $V_{\tilde{x}}(\infty)/V_{\tilde{x}}(0)$ increases to 0.5 when the parameters of the selected model

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* It is demonstrated in [5] that when m=40 the error in calculating the dispersion by formula (8) does not exceed 5%.

** In Fig. 1 b and c: τ_{cor} and V_v -- parameters of selected statistical model; τ_{corr} and V_{av} -- corresponding actual quantities; $V_{\tilde{x}}(\infty)$ -- estimate error dispersion in steady-state, when $V_{\tilde{x}}(k)=V_{\tilde{x}}(k-1)$.

deviate from the actual situation for the algorithm not allowing for cross-cumulants, in the following proportions: the observation noise level of the model V_v becomes 10.5 times greater, or 8 times smaller than the actual value V_{vp}, while the correlation coefficient τ_{cor} becomes w times greater, or 10 times smaller than τ_{corr}. V_v in the algorithm which allows for cross-cumulants may be 25 greater or 12 times smaller than V_{vp}, and τ_{cor} can be 7 times greater or 12 times smaller than τ_{corr}. When there are large deviations of the parameters of selected models from the actual situation the algorithm diverges (the a posteriori dispersion exceeds the a priori $V_{\tilde{x}}(\infty)>V_{\tilde{x}}(0)$). In the algorithm which does not allow for cross-cumulants the divergence begins when V_v is 90 times greater or 75 times smaller than V_{vp}, and when τ_{cor} is 20 times greater or 80 times smaller than τ_{corr}. In the algorithm which allows for cross-cumulants this situation occurs when V_v is 130 times greater or 80 times smaller than V_{vp}, and when τ_{cor} is 30 times greater or 90 times smaller than τ_{corr}. When the system dimensionality is greater, the sensitivity of algorithms which allow for cross-cumulants is still smaller.

It is apparent from the figures above that algorithms which allow for cross-cumulants provide better estimation accuracy and low sensitivity of deviation of the parameters of the selected model from the actual situation. However, they are significantly more complex than algorithms which do not allow for cross-cumulants, which limits their use.

Let us now estimate the difficulty of implementing algorithms for processing multi-dimensional signals and interference. In most cases these algorithms must be carried out in real time on special processors. Let us determine the speed required of this processor. In order to do this, we must count up the total number of arithmetic operations executed in one iteration of the algorithm. The total number of additions and subtractions N_1, as well as multiplications and divisions N_2, which are executed in one iteration of an algorithm which allows for cross-cumulants can be found from the following equations:

$$N_1 = 5n^3 - 2n^2 + 3n^2m + 2nm^2 - nm - n, \tag{9}$$

$$N_2 = 5n^3 + n^2 + 3n^2m + 2nm^2 + 2nm. \tag{10}$$

For the algorithm which does not allow for cross-cumulants the number of additions and subtractions N_3 and multiplications and divisions N_4 are found from:

$$N_3 = n^2 + 3n + 2m, \tag{11}$$

$$N_4 = 6n^2 + 7nm. \tag{12}$$

The quantities N_1, N_2, N_3, N_4 are found by analyzing formulas (3)-(7). If n=m, expressions (9)-(12) reduce to the form:

$$N_1 = 10n^3 - 3n^2 - n, \quad (9') \qquad N_2 = 10n^3 + 3n^2, \quad (10')$$

$$N_3 = n^2 + 5n, \quad (11') \qquad N_4 = 13n^2. \quad (12')$$

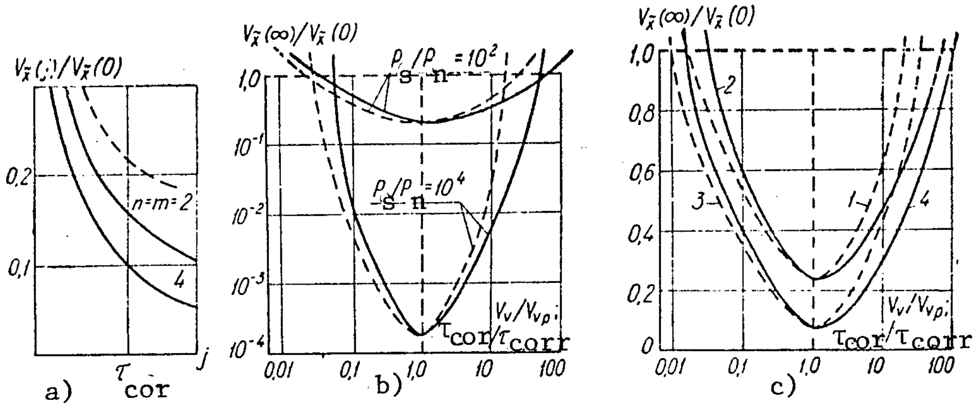

Fig. 1

For n=2, we obtain N_1=66, N_2=92, N_3=14, N_4=52. For n=4 -- N_1=588, N_2=688, N_3=86, N_4=208. When the algorithm which does not allow for cross-cumulants is implemented on a processor, 4.7 times fewer additions and subtractions and 1.78 fewer multiplications and divisions must be done for the two-dimensional case than for an algorithm which allows for cross-cumulants. These figures for the four-dimensional case are 16.3 and 3.3.

Our analysis of algorithms for processing multidimensional signals and interference indicates that algorithms which allow for cross-cumulants have advantages in estimation accuracy and sensitivity to deviation of model

parameters from the actual signal-interference situation over algorithms which do not allow for cross-cumulants. However, the algorithms which allow for cross-cumulants are significantly more complex than algorithms which do not, and require the use of faster and more expensive processors. Therefore, algorithms for processing multidimensional signals and interference which allow for cross-cumulants should be used when high estimation accuracy is required and which cannot be provided by algorithms which do not allow for cross-cumulants, or when the parameters of the selected model may diverge significantly from the actual situation. Otherwise it is best to use simpler algorithms which do not allow for cross-cumulants.

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ADAPTIVE COMPENSATION FOR QUASI-DETERMINATE NARROWBAND INTERFERENCE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian
Vol 26, No 4, Apr 83 (manuscript received 1 Nov 82) pp 103-104

[Article by V.G. Tatarnikov]

[Text] A requirement for suppressing narrowband interference often accompanies the solution of problems involved in receiving and processing radio signals [1]. Frequency, time and space selection methods are commonly used for this; however, when the interference is quasi-determinate, direct compensation may be more effective [2].

Since they are characterized by substantial interference suppression (by 40 dB or more), compensation systems provide almost a complete absence of distortions of the valid signal, which is an undisputed advantage over other systems; their use can, in part, improve the electromagnetic compatibility of radio equipment. Indeterminacy of the parameters of the interference can be overcome by using adaptive signal processing algorithms.

Let us assume that the input to the receiver is the additive mixture $y(t)=$ $=s(t)+n(t)$, where $s(t)$ -- valid signal, being a random process with spectral density approximately constant in the vicinity of frequency ω_0 and equal to S_0;

$$n(t) = U(t)\{M_c \cos[\omega_0 t + \varphi(t)] - M_s \sin[\omega_0 t + \varphi(t)]\} \quad (1)$$

-- quasi-determinate narrowband interference with random parameters M_c and M_s; the quantities $U(t)$, ω_0 and $\phi(t)$ are assumed to be known. One example is a situation in which the interference source is a radio transmitter whose signal may reach the observer over a channel with a high degree of protection against outside influences (e.g., cable); the occurrence of interference at the input of the receiver is caused by factors which cannot, and should not, be accounted for precisely (static induction, reflections from nearby objects, etc.). The observer has available the process $n_0(t)=$ $U(t)\cos[\omega_0 t \pm \phi(t)]$; the noise $n(t)$ is the copy $n_0(t)$ with changed scale and shifted phase, with the amount of change in scale and phase unknown.

The problem of compensating for interference n(t) is equivalent to that of determining the unknown quantities M_c and M_s (and when M_c and M_s are known, the problem reduces to simple subtraction of two signals). The estimated values (M_c^* and M_s^*, respectively), obtained in some way, should be used instead of the actual unknown values of M_c and M_s. In particular, we can use optimal mean square estimates which are solutions of the system of Kalman-Busey equations from [1], which in the present case reduce to the form

$$dM_c^*/dt = 2k(t)(y_t - n_t^*)\cos xU(t)/S_0, \tag{2}$$

$$dM_s^*/dt = -2k(t)(y_t - n_t^*)\sin xU(t)/S_0, \tag{3}$$

where n*(t) -- estimated value of noise obtained from expression (1) by replacing M_c and M_s with M^*_c and M^*_s, respectively; $x=\omega_0 t+\phi(t)$; k(t) -- dispersion of deviation of error, equaling

$$k(t) = S_0 k_0 \Big/ \left[S_0 + k_0 \int_0^t U^2(t)\,dt \right]. \tag{4}$$

Here k_0 -- a posteriori dispersion of the quantities M_c and M_s. Equations (2) and (3) are valid if the correlation time of s(t) is negligibly small as compared with the system adaptation time. The dispersion of the deviation of estimate drops monotonically with time, approaching zero without limit; consequently, arbitrarily accurate interference compensation is theoretically possible given a long enough time interval.

Equations (2) and (3) are rather complicated for hardware implementation. Considerable simplification can be achieved by replacing $k(t)U(t)/S_0$ with μ=const; then we obtain

$$dM_s^*/dt = -2\mu(y_t - n_t^*)\sin x. \tag{5}$$

$$dM_c^*/dt = 2\mu(y_t - n_t^*)\cos x; \tag{6}$$

We note that the familiar Widrow-Hoff algorithm [2] follows from the latter equations with μ replaced by $\nu U(t)$ and going from continuous to discrete time.

It can be demonstrated that the deviations of the estimates $\Delta M_c = M_c - M^*_c$ and $\Delta M_s = M_s - M^*_s$ obtained in solving equations (5) and (6) are, when the spectrum of the valid signal is broad enough, exponentially correlated Markov processes with dispersion $\sigma^2 = \mu U^2 S_0/4$ and spectrum cut-off frequency $\omega_{\Delta M} = \mu U$.

In contrast to the optimal algorithm, the average power of the estimation error (5), (6) does not decrease with time; rather, it is a quantity proportional to the interference power. Another deficiency of the quasi-optimal algorithm is the significantly longer adaptation time. Analysis indicates that the average adaptation time, defined as the time during which the absolute deviation of the estimate drops from the initial value of $\sqrt{k_0}$ to some small value ε, is

$$t_{\varepsilon\,opt} = S_0/U^2\varepsilon, \tag{7}$$

for the optimal algorithm, and

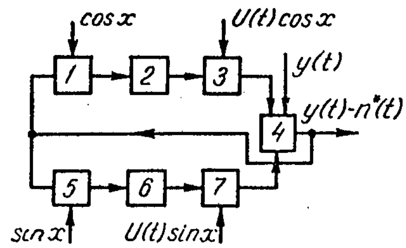
$$t_\varepsilon = \ln(\sqrt{k_0}/\varepsilon)/\mu U \tag{8}$$

for the quasi-optimal one, and for the values entering into these expressions which are encountered in practice the radio $t_\varepsilon/t_{\varepsilon opt}$ lies within the limits 10-10^6. Furthermore, an interference compensation system operating in accordance with algorithm (5), (6) suppresses the low frequency components of the valid signal; therefore, the quantity μ must be chosen to satisfy the condition that the time constant of the system, equaling $1/\mu U$, is sufficiently large.

In spite of the deficiencies enumerated above, the quasi-optimal system is preferable in most cases because of the simplicity of hardware implementation, as well as the capability of adapting to slow fluctuations in the characteristics of the components (e.g., drift in the output voltage of the integrators, etc.).

Figure 1 shows the structural diagram of a system which implements algorithm (5), (6), where: 1,3,5,7 -- multipliers; 2,6 -- integrators; 4 -- adder.

Fig. 1



In order to verify the theoretical results experimentally a laboratory model of the quasi-optimal interference compensation system was built. During the design and fabrication of the model, the basic requirements for the parameters of the system components were determined. It was established in part, that small phase shifts (of up to 5-10°) of the reference signals with respect to the values indicated in the figure, as well as differences in their amplitude, do not have a significant effect on compensation accuracy. Also of minor significance are nonlinearity of the relationship between the gain of modulators 3,7 and the control voltage, and nonlinearity of the relationship between the output voltage of the phase-sensitive detectors 1,5 and the amplitude of the input signal. Conversely, compensation accuracy is strongly influenced by the amount of zero drift of multipliers 1,5 and the operational amplifiers making up the integrators, and, for U(t)≠const, nonlinear and frequency distortions of the envelope of radio signals at the output of modulators 3,7. For example, when zero drift occurs in multiplier 1 the system responds with a corresponding increase in the cosine component of n*(t), which is equivalent to an increase in the uncompensated residue and a sharp drop in compensation accuracy. Practical investigation of different types of components showed that the best results are provided by using phase-sensitive detectors with field-effect transistors operating in the unsaturated mode in the capacity of multipliers 1,5, and using balanced modulators with resistor optrons in the capacity of multipliers 3,7. These detectors have minimal zero drift, and the decoupling between the inputs is on the order of 60 dB. Resistor-optron multipliers, each representing a pair of variable attenuators connected in a differential-bridge arrangement, provide minimum distortion of the envelope with acceptable amplitude-phase conversion values.

This model provided suppression of interference with carrier frequency of 31.8 MHz and spectrum width at the -40 dB level of the order of 200 kHz to a level of at least -46 dB with respect to the input. The actual value of the interference voltage at the input varied within limits of 3-25 mV, and the phase varied from 0 to 2π. The adaptation time constant was 0.5 sec.

It can thus be concluded that the use of the present system provides significant attenuation of quasi-determinate narrowband interference that is not accompanied by distortion of the valid signal.

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CSO: 8144/1394

UDC 621.315.2

INTERFERENCE VOLTAGE INDUCED BY EXTERNAL SOURCES IN MULTICONDUCTOR COMMUNICATION LINES

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83
(manuscript received 5 May 82) pp 41-44

SHKARIN, Yu. P.

[Abstract] The interference voltage which external sources induce in a multiconductor communication line is calculated on the basis of the telegraphist's equations for a homogeneous n-conductor line with distributed voltage and current sources. Expressions are obtained for the interference voltage at the beginning and at the end of such a line, taking into account reflections at both ends. The general results are applied to the specific case of a 3-conductor 500 kV overhead transmission line near a radio station operating at 300 kHz. All three conductors (phase) are in one horizontal plane, spaced symmetrically with a 10 m distance between the center conductor and each outer one. The load is connected either across an outer phase and ground or across an outer phase and the center phase. The interference level is 24.5 dB higher in the first case. References: 4 Russian.
[232-2415]

UDC 621.376.56

PARAMETERS OF QUANTIZER IN RADIO COMMUNICATION SYSTEM WITH PULSE CODE MODULATION

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 16 Sep 82) pp 44-48

VELICHKIN, A. I., SUROVTSEV, Yu. A. and TATARSKIY, B. G.

[Abstract] A quantizer of a continuously significant message can be characterized by two parameters: level quantization step and mean-square quantization error. An efficient approximate of calculating the mean-square

error for a message with many quantization levels is representing the mean-square error as a sum of integrals, one for each quantization step, and expanding the probability density of the message in each into a Taylor series at the center of the integration interval. The method is applied, with appropriate refinements, to unipolar messages with various probability distributions (Rayleigh, exponential, bell-curve) and to symmetric messages with various probability distributions (normal, exponential). Formulas and numerical data are shown. Tables 2; references 2: 1 Russian, 1 Western.
[226-2415]

UDC 621.391.019.4

ESTIMATION OF NOISE PARAMETERS IN DETECTORS OF OPTICAL SIGNALS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 4 Oct 82) pp 48-52

TOLPAREV, R. G. and BORISOV, E. V.

[Abstract] Estimating the noise parameters in a detector of optical signals is not easy in the case of an insufficiently large sample volume and consequently short stationarity intervals. An expedient method of estimating them under such conditions is based on representation of the random process at the output of the noise estimation channel (counter of noise photoelectrons) within a j-th locally-stationary interval as a vector of independent readings $K = (k_1, k_2, \ldots, k_N)$ with some conditional distribution $P(k_i/s_{nj})$. A wide range of practical situations is covered by assuming this distribution to be a negative-binomial one and the a priori probability density distribution $w(s_{nj})$ to be a log normal one. The mean number of noise photoelectrons s_{nj} can be easily estimated on this basis. This is demonstrated on such a case, with the a priori probability density distribution having only one parameter: dispersion σ^2. Formulas are based on these and other distributions $P(k_i/s_{nj})$ (geometric, gamma, Poisson, normal) and $w(s_{nj})$ (m-, uniform) in all possible pairwise combinations. Tables 1; references 5: 4 Russian, 1 Western (in translation).
[226-2415]

UDC 621.391.1.019.4

RECEPTION OF DIGITAL SIGNALS APPEARING WITH PULSE INTERFERENCE

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 8 Oct 82) pp 8-12

YERSHOV, L. A. and IVANOV, A. V.

[Abstract] An optimum receiver is synthesized for discrete signals appearing with both white noise and random pulse interference. An additive mixture $\xi(t) = \mu S_1(t) + (1 - \mu)S_2(t) + n(t) + \gamma(t)$ is assumed to appear at the receiver input during the observation time interval $[0,T]$, where $S_1(t)$ and $S_2(t)$ are two deterministic signals, μ is a random parameter equal to 0 or 1 with equal probabilities $p(0) = p(1) = 0{,}5$, $n(t)$ is a Gaussian white noise, and

$$\gamma(t) = \sum_{i=0}^{k} A_i f(t - t_i)\cos(\omega_0 t + \varphi_i)$$ is a random sequence of nonoverlapping interference radio pulses of duration $\tau_p \ll T$ with mutually independent random amplitudes A_i and initial phases φ_i. For optimal discrimination of signals $S_1(t)$ and $S_2(t)$, the hypotheses of $\mu = 0$ or $\mu = 1$ are tested on the basis of the likelihood ratio according to statistical decision theory. The receiver structure executing this algorithm contains two channels partially compensating the effect of pulse interference. The interference immunity of such a receiver is compared with that of a correlational receiver for wideband noiselike signals, the latter operating with M-sequences, when both receive phase-keying signals under the same conditions of white noise and pulse interference. The results for M-sequences such as $B_1(x) = x^7 + x^3 + 1$ and $B_2(x) = x^6 + x + 1$ indicate that a correlational receiver is preferable when the interference-to-signal ratio is $q < 10$ and a compensating receiver is preferable when this ratio is $q > 10$. Figures 2; references: 5 Russian. [226-2415]

UDC 621.391.828;621.396.677.53(088.8)

COMPONENT SELECTION OF ELECTROMAGNETIC SIGNALS

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 11 Dec 81) pp 72-73

GAVRILIN, A. T., GRECHIKHIN, A. I. and TOROPOV, L. A.

[Abstract] Separation of two radio signals from different sources or of a useful signal from an interference signal within the same frequency band and with the same polarization, and propagating in the same direction, is possible by a method which involves component selection in the electromagnetic field. This method is based on the departure, under certain conditions, of the ratio of complex amplitudes E/H (electric field intensity E, magnetic field intensity H) from $Z_0 = 120\,\pi$ ohms. The difference can be appreciable when the receiver is located within the induction zone of both signal sources (respective distances $R_{1,2} \leq \lambda/2\pi$ with $R_1 \neq R_2$ or with

$R_1 = R_2$ but sources of different kinds) or within the induction zone of one ($R_1 \leq \lambda/2\pi$) and within the wave zone of the other ($R_2 >> \lambda/2\pi$), or when the signals propagate through a bounded space such as the gap between earth and ionosphere. The method is not only applicable where conventional methods are inadequate or not feasible but is also reversible, i.e., an antenna built on this principle can be used for field suppression and interference-free transmission. Where other methods of signal separation can be used combining them with this method will either simplify the process or improve the results. References: 6 Russian.
[225-2415]

UDC 621.395.5:621.372.544

DIGITAL FILTERS FOR CONVERSION OF ANALOG SIGNAL TO PULSE-CODE-MODULATION SIGNAL

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83
(manuscript received 27 Nov 81) pp 49-52

GOL'DENBERG, L. M., BRUNCHENKO, A. V., MATYUSHKIN, B. D., POLYAK, M. N., TALANOV, A. O. and YAKOVLEV, L. A.

[Abstract] An excellent method of converting an analog signal to a PCM signal for digital data transmission is converting it to a delta-modulation signal first and then the delta modulation signal to a PCM signal with subsequent digital filtration. Here the problem of digital filtration in such a system is considered from the standpoints of performance and design. On the basis of known signal and noise characteristics, the maximum noise reduction index attainable with an "ideal" (distortionless and noiseless) digital filter is calculated in terms of the ratio of noise power in the analog signal to noise power in the PCM filter output signal. An optimum digital filter within a given class is then synthesized according to the criterion of maximum noise reducgion at given cutoff and discretization frequencies with given allowable distortion of the amplitude-frequency characteristic. The feasibility of reducing the noise power by means of simple digital filters, specifically a nonrecursive one with the transfer function $H_n(z) = d_1 + d_2z^{-1}$ and a recursive one with the transfer function $H_r(z) = r_1/1 = r_2z^{-1})$, is examined without an a priori constraint of a linear phase-frequency characteristic. Calculations involve first normalizing the generally nonlinear amplitude-frequency characteristic and then solving the nonlinear algebraic equations for filter operation in a binary code. A typical numerical example demonstrates the practicality of digiral filtration in analog-to-PCM conversion systems. Figures 4; tables 2; references 4: 1 Russian, 3 Western (1 in translation).
[232-2415]

UDC 621.395.345

EXPERIMENTAL OPERATION OF QUASI-ELECTRONIC AUTOMATION INTERURBAN TELEPHONE STATION

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83
(manuscript received 18 Nov 81) pp 7-13

KOBLENTS, Ya. G., YAKOVENKO, D. A. [deceased], and KIRSANOV, V. I.

[Abstract] The "Kvarts" quasi-electronic automatic interurban telephone station which was installed in Leningrad in July 1980 after completion of line tests includes a 1024 x 2 input/output module and a 6-stage switching system with 64 x 64 2-stage input and output groups, each containing sixteen 8 x 8 x 4 ferreed (reed relay on ferrite) matrix connectors. Data exchange between the switching system and telephone sets on the one hand and a "Neva-1" special-purpose oontrol and computation complex on the other is facilitated by a peripheral control device. Experimental operation was begun in 1980 and carried out during the 8:30 AM to 6:00 PM hours in the August-December period, with institutional subscribers loading the lines. The special-purpose control and computation complex was organized in two modes, first with a semiautomatic 2-machine set and then with a 1-machine simplex for algorithm and program debugging. Both hardware and software were tested for performance and reliability. An equipment defectiveness and failure analysis on the basis of the test data revealed that press-in plugs are by far the least reliable components. They will be replaced with cut-in plugs. The software for the experimental operation included 296 kbyte basic operational programs, 90 kbyte technological telephone programs, 50 kbyte operational control programs, and a 74 kbyte starting-training program with various starting-debugging diagnostic programs. Subsequent operation of the "Kvarts" station during the 1981-82 period confirmed the results of the experiment. Figures 9; tables 1.
[232-2415]

UDC 621.395.345

REFINEMENT OF RELIABILITY INDICATOR SYSTEM FOR SWITCHING JUNCTIONS AND STATIONS

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83
(manuscript received 3 Feb 82) pp 30-32

ZELENTSOV, B. P. and SUTORIKHIN, N. B.

[Abstract] The possibility of failure to detect a fault in a functional module is considered, such a situation occurring for any of at least three reasons or any combination thereof: 1) Faults are checked only during periodic inspections; 2) Faults are not detected by a periodic-inspection system and remain undetected either until the next inspection or until an

inspection by another system; 3) Faults are checked by a continuous-inspection system which admits Type-2 errors ("missed hit"). On the other hand, a perfect functional module can be erroneously interlocked because of a Type-1 error ("false alarm") in the inspection system. These situations are covered in a reliability analysis of switching systems. The reliability indicators of such systems are refined by taking into account the reliability of its functional modules as well as of the inspection equipment. These indicators include mean time between interlocking events, mean time of remaining interlocked, mean time of correct operation, mean time of incorrect operation with undetected fault, percent idle time in interlocked state, nominal readiness and idleness factors. References: 7 Russian.
[232-2415]

UDC 621.395.374

TERMINAL AUTOMATIC INTERURBAN TELEPHONE STATION USING ARM-20 AND ARYe-13 EQUIPMENT

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83 (manuscript received 9 Mar 82) pp 13-17

GERCHIKOV, Ye. Ya. and PETEL'SKIY, P. V.

[Abstract] A terminal automatic interurban telephone station is described which uses basic ARM-20 coordinate equipment and its modernized ARYe-13 version with relay control replaced by electronic control. The station structure consists of a switching system and a control system. The control system includes four input-output devices, a data processing module, and an interface from the latter to the switching system. The data processing module contains up to 15 TRS call processors and 1 OMR operational data processor, both types using the same hardware but different software, a TsZU central memory and an MUX multiplier. The station software includes control programs and equipment operation programs as well as call processing programs. The power supply is provided by a 48 V set of storage batteries with trickle charge through a thyristor rectifier bank. Provisions have been made for combined operation of ARM-20 and ARYe-13 components. Counters have been installed in the ARYe-13 system for computing the cost of telephone conversations. Figures 4; references: 2 Russian.
[232-2415]

UDC 621.396.4

HETERODYNE WITH RECEIVER-TRANSMITTER FREQUENCY SHIFT IN INTERMEDIATE RADIO RELAY STATION

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received, after revision, 16 Apr 82) pp 86-87

PROTOPOPOV, A. P., SHEVCHENKO, V. I. and CHEREPUKHIN, V. I.

[Abstract] A heterodyne for an intermediate radio relay station is described which ope ates with a receiver-transmitter frequency shift. It contains two microwave oscillators, one for the receiver mixer and one for the transmitter mixer, and a reference quartz oscillator which stabilizes the frequency difference. The effect of frequency drift in both receiver and transmitter is eliminated, the instability of the relay output frequency being determined solely by the instability of the frequency shift: $\Delta f_{trans,out}/f = \Delta f_{rec,in}/f - \Delta f_{sh}/f$. Instability of the receiver frequency is minimized by insertion of a high-Q dielectric resonator holding it within 10^{-4} over the $\pm 50°C$ temperature range. The two microwave signals, from the heterodyne receiver and from the heterodyne transmitter, combine in a shift mixer. The difference-frequency output signal of the latter passes through a pulse-type frequency divider to a pulse-phase detector. The output signal of the reference quartz oscillator passes through another pulse-type frequency divider to the same pulse-phase detector. Here the phases of the two signals are compared by the method of voltage sampling and storage, ensuring suppression (to -80 dB or less) of the mismatch over a 360° range of phase shift. Both heterodyne oscillators are built with Schottky-barrier microwave transistors in the common-source connection. Figures 3; references: 2 Russian.
[225-2415]

UDC 621.396.14

POWER AMPLIFICATION OF COMPLEX SIGNALS WITH PARALLEL STRUCTURE

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received, after completion, 20 Apr 82) pp 56-57

GOLIKOV, O. B. and YAKOVLEV, A. I.

[Abstract] The feasibility of using the peak power of a transmitter for amplification of complex signals with parallel structure is established on the example of such a signal consisting of N harmonic segments with equal amplitudes and orthogonal throughout the signal duration. In the worst case of a constant phase code the ratio of peak signal power to average signal power is $\gamma = \sqrt{2N}$ and thus the power gain is N times lower than in the worst

case of a sequential signal structure. Considering that the distribution of instantaneous signal values tends to become normal as the number of harmonic segments increases (N > 10) with random or pseudorandom distribution of initial phases, the only possible but very effective method of increasing the gain to levels attainable for signals with sequential structure is separate amplification of each signal element and subsequent addition of all across the terminal load. Figures 2; references: 4 Russian.
[225-2415]

DEVELOPMENT AND INSTALLATION OF SUSPENDED AND BUILT-IN COMMUNICATION CABLES

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83 (manuscript received 4 Aug 80) pp 45-48

KALYUZHNYY, V. F.

[Abstract] Various communication cables recently developed for installation along electric transmission lines, either self-supporting or built-in, were displayed by AEG-Telefunken and Felten-Guilleaume at the "Svyaz'-81" International Exhibition. They include the self-supporting symmetric A-SLH2Y2Y1x4x0.9 and coaxial A-SLH2YLb2.3/10, and the built-in coaxial A-PSLH6Ybb1x0.9/3.5, all three manufactured by AEG-Telefunken and having excellent attenuation characteristics. Isolation transformers for connecting tone-frequency channels and multichannel transmission lines to self-supporting or built-in symmetric or coaxial communication cables are now manufactured by the Yugoslav firm "Iskra". In England and other western countries optical cables are suspended on the supports of electric transmission lines, a typical example being a loop of optical fibers to hang on supports of a 400 kV transmission line. Figures 6; tables 2; references 10: 1 Yugoslav, 9 Western.
[232-2415]

NEW ENGINEERING DECISIONS IN AREA OF HIGH-FREQUENCY COMMUNICATIONS OVER SUPERHIGH VOLTAGE ELECTRIC POWER TRANSMISSION LINES

Moscow ENERGETIK in Russian No 5, May 83 pp 22-23

ISHKIN, V. Kh., candidate of technical sciences and TSITVER, I. I., engineer

[Abstract] A study is made of the technical solutions of two items related to the problem of electromagnetic compatibility of HF channels in a distributed electric power network within a strictly regulated frequency band, and the creation of HF channels for multichannel telephone communications over superhigh voltage electric power transmission lines. New types of HF lines have been introduced for multichannel telephone information transmission systems which are utilized for 3- and 12-channel communications systems. They operate betwe n current conducting lines, within phases or within lines. HF communications channels can also be organized using lightning protection conductors. Because of the short distance between individual conductors

of a single phase, systems operating within one line are naturally well-balanced, yielding comparatively little transient attenuation in other channels organized over the same or neighboring power transmission lines and very low levels of HF interference. Identical frequencies carrying two different messages cannot be used over the same electric power transmission line. Figures 2.
[244-6508]

UDC 621.391.25

METHOD OF SYNTHESIZING BALANCED ALPHABETIC CODES FOR DIGITAL TRANSMISSION SYSTEMS

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received, after revision, 14 Jun 82) pp 65-66

MARKARYAN, G. S.

[Abstract] A synthesizer of alphabetic multiposition codes has been developed for more efficient operation of digital transmission systems with linear multilevel signals. After optimization of the matrix of mean values, dispersions, and transition probabilities, the coder (finite discrete automaton) generates codes which are without a constant component and thus balanced and have only a small low-frequency component, the object being either to reduce the cycle frequency of the linear signal and thus increase the word length or increase the transmission rate without change of the cycle frequency. Examples of such codes are 8B-6T, 6B-4T, and 12B-6Q reducing the cycle frequency by a factor of 1.33, 1.5, and 2.0 respectively. A comparison of the normalized energy spectrum of the 8B-6T code with those of the known 4B-3T and MS 43 codes reveals a smaller low-frequency component here, with a 1.99 mean error multiplication factor. In the proposed codes there is a higher probability of appearance of significant symbols, which facilitates extraction of timing information. These codes also provide the possibility of checking the reliability of transmission. Figures 2; references 3: 2 Russian, 1 Western in translation.
[225-2415]

UDC 621.391.272:621.396.237

SIGNAL DISTORTION IN COMMUNICATION CHANNELS WITH LARGE-SCALE DISCRETE SHIELDING INHOMOGENEITIES

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 11 Dec 82) pp 13-15

PONOMAREV, G. A. and FORTES, V. B.

[Abstract] Transmission of arbitrarily modulated radio or television signals through a linear channel with large partial rereflectors such as urban structures is considered, assuming a normal distribution of both orghogonal components of its complex transfer function RE $H(\omega,r,t)$, Im $H(\omega,r,t)$ with zero mean and a uniform distribution of the phase over the $[0,2\pi]$ interval. The frequency-selective distortion and fadeout of signals in such a channel are evaluated on the basis of the statistical model of a channel with discrete shielding inhomogeneities. The elements of the corresponding correlation matrix are determined in terms of the delay energy spectrum, which characterizes the distribution of mean energy over signal beams arriving at the receiver. The behavior of the frequency correlation of the channel is analyzed generally, with the aid of Fourier transformation. The envelope of the output signal is calculated specifically for a simple square-pulse input signal of duration T. The two extreme cases of very long and very short duration are of particular interest. The envelope has similar shapes, except that the fall time is long and there is no flat top in the latter case. Figures 1; references 6: 3 Russian, 3 Western.
[226-2415]

UDC 621.375.9:621.391.268

ULTIMATE SENSITIVITY OF CORRELATIONAL AMPLIFIER WITH COMPENSATION

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received, after revision, 25 May 82) pp 60-62

CHEREVKO, A. G.

[Abstract] A correlational amplifier with compensation and with an integrating output stage has been proposed (USSR patent disclosure No 298899, 1971) for measuring the dispersion of noise signals by the null method. In most practical cases the input voltage and the compensating voltage are each a white noise, with a noise diode whose spectral noise density depends linearly on the anode current serving as a source of the compensating voltage. The ultimate sensitivity of such an amplifier is calculated for this case, assuming both processes to be normal ergodic ones wth constant spectral density within the amplifier passband. The results are useful for estimating the random error in measurement of fluctuation processes with a signal weaker than the intrinsic instrument noise. Figures 1; references 6: 4 Russian, 2 Western.
[225-2415]

UDC 621.385.69

MILLIMETER-WAVE AMPLIFIER BASED ON ELECTRODYNAMIC STRUCTURE OF DIFFRACTION RADIATION GENERATOR

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 15 Feb 82) pp 93-94

KORNEYENKOV, V. K, and MIROSHNICHENKO, V. S.

[Abstract] The electrodynamic structure of a diffraction radiation generator can be used as a low-power low-noise amplifier for the millimetric wave band. The feasibility of attaining gain levels of 20-30 dB has been

demonstrated both theoretically and experimentally. The mechanism of signal amplification in such a structure consisting of an open resonator with a periodic grating on one of the mirrors is based on excitation of a natural mode and a field of space harmonics by the signal from an external source. An electron beam passing by the structure at a velocity nearly equal to the phase velocity of any one harmonic will first be bunched, whereupon its energy will be emitted into the resonator space in the form of diffraction radiation. As long as the beam current remains below the generator starting current, the device operates as a regenerative amplifier with a low noise level ensured by the high selectivity of the high-Q resonators as well as by feedback on the fast wave equivalent to directional discrimination of radiation and facilitating suppression of random beam fluctuations. The noise level is also reduced by placing the amplifier input and output stages on the opposite mirror, far from the electron beam. The power gain then depends on the ratio of beam current to generator starting current $I/I_s < 1$ and on the resonator-load coupling coefficient. As $I \to I_s$, the gain should increase and the bandwidth should decrease. This has been confirmed experimentally. The measured amplitude characteristic fits the theoretical relation for power gain

$$K_p = \left[\frac{R_c + (I/I_s)^{3/2}}{1 - (I/I_s)^{3/2}}\right]^2$$

(R_c - reflection coefficient in "cold" resonator, ranging from -1 to 1). With the reflection coefficient in the $0 \leq R_c \leq 1$ range, the gain increases monotonically with the current and exceeds 20 dB within the self-excitation range $0.9 < I/I_s < 1$. Within this range, however, the gain begins to decrease sharply as the output power exceeds 0.1 mW. Typical performance characteristics at $I/I_s = 0.9$ are power gain K_p=20 dB, maximum P_{out}=200 mW, and bandwidth Δf=15 MHz. Figures 2; references: 4 Russian.
[225-2415]

UDC 517.432.1

USE OF FAST FOURIER TRANSFORMATION IN SIMULATION PROBLEMS

Kiev ELEKTRONNOYE MODELIROVANIYE in Russian Vol 5, No 3, May-Jun 83 (manuscript received 27 Mar 81, after completion 3 Aug 81) pp 94-98

SHILOV, ALEKSANDR MIKHAYLOVICH, candidate of technical sciences, head of laboratory, Tallin Electrical Engineering Plant

[Abstract] The method of fast Fourier transformation can be used for performing an inverse Fourier integral transformation, generally to function f(t) in time $0 \leq t < \infty$ from its tranform in the frequency domain. The integral is evaluated according to the trapezoidal rule, upon replacement of the infinite limit with aafinite one and subsequent evaluation of the error, and use is made of the Euler relation $e^{\pm j\omega t} = \cos\omega t \pm j\sin\omega t$. The algorithm has been constructed and the cosh function has been tabulated for a computer, discretization of the frequency domain into $N - 1 = 2047$ intervals being most efficient and adequatley accurate. The practical application of this method is demonstrated on simulation of transient processes in semiconductor devices such as a thyristor, which involves solving the equations of kinetics of excess electron and hole concentrations in the base region, in the one-dimensional approximation and assuming a low injection level. Tables 2; references: 2 Russian.
[238-2415]

UDC 681.34:536.21

SPECIAL-PURPOSE HYBRID COMPUTER COMPLEX FOR SIMULATION OF PHYSICAL FIELDS

Kiev ELEKTRONNOYE MODELIROVANIYE in Russian Vol 5, No 3, May-Jun 83 (manuscript received 23 Nov 81) pp 44-47

KUNESH, IOZEF [KUNEŠ, J.], candidate of technical sciences, head of department, Central Scientific-Research Institute of Concern "Shkoda", Pl'zen', ChSSR; BARTIK, YAN [BARTIK, J.], engineer, Central Scientific-Research Institute of Concern "Shkoda", Pl'zen', ChSSR; FRANTA, VATSLAV, candidate of technical sciences, head of laboratory, Central Scientific-Research Institute of Concern""Shkoda", Pl'zen', ChSSR; and VAVROKH, OTAKAR, senior engineer, Central Scientific-Research Institute of Concern "Shkoda", Pl'zen', ChSSR

[Abstract] The SKODA JHS-1 computer has been developed for simulation of physical processes and fields as well as for solution of reverse and optimization problems in all branches of engineering. It is a hybrid system consisting of an analog processor interfaced with a measuring module and a control module through data input/output devices and a memory. Three printers are connected to it through an analog-to-digital converter and a recording instrument is connected directly to the memory. The analog processor is a standard general-purpose bank of RC networks. The measuring subsystem is the key component, measurements being made by time quantization of analog signals with the aid of a function generator, a step counter, control signals and time delay. Both analog and digital memories are contained in a single module. Small size and high reliability have been attained through maximum possible circuit integration or hybridization. The complex can be coupled to a separate digital computer or microcomputer. Figures 8; references 10: 3 Russian, 2 Czechoslovak, 5 Western.
[238-2415]

UDC 621.311.019.3:681.323

SIMULATION OF DEVELOPING FAULT PROCESSES ON PARALLEL-SERIES COMPUTER STRUCTURES

Kiev ELEKTRONNOYE MODELIROVANIYE in Russian Vol 5, No 3, May-Jun 83 (manuscript received 25 Dec 81) pp 21-26

VLAZNEV, IGOR' KONSTANTINOVICH, chief engineer, Institute of Modeling of Power Engineering Problems, UkrSSR Academy of Sciences, Kiev; GORSKIY, YURIY MIKHAYLOVICH, candidate of technical sciences, senior scientific research worker, Siberian Power Engineering Institute, Siberian branch, USSR Academy of Science, Urkutsk; DODONOV, ALEKSANDR GEORGIYEVICH, doctor of technical sciences; KASHTANOV, YURIY BORISOVICH, junior research worker, Siberian Power Institute, Siberian Department, USSR Academy of Sciences, Irkutsk; FEDOTOV, NIKOLAY VASIL'YEVICH, candidate of technical sciences, senior scientific research worker, Institute of Problems of Modeling in Power Engineering, UkrSSR Academy of Sciences, Kiev; SHCHETININ, ALEKSANDR MIKHAYLOVICH, candidate of technical sciences, senior scientific research worker, Institute of Problems of Modelingiin Power Engineering, UkrSSR Academy of Sciences, Kiev

[Abstract] An apparatus is developed for simulation and diagnosis of fault processes, typically those occurring in electric power networks. Simulation in real time requires a speculative model of the object which will reveal its operation and transitions from one state to another. Diagnosis involves identification of the present state and analysis of the process dynamics. Diagnosis can be followed by control for purposes of stabilization, recovery, and failure prevention. The problem is treated according to the theory of graphs and the theory of finite-automata, a class-1 or class-2 reverse one being most suitable, whereupon a parallel structure is synthesized for computing extremal and optimal paths from initial to final points. The structure consists of a minicomputer and a special-purpose digital analog interfaced with it, the latter essentially being a multiprocessor. Here the graph is constructed by a series of arc modeling processors, all combined into a single module with inputs and outputs automatically switched according to the graph topology. The operation of this digital analog and its components is synchronized by a control module. The computer structure can predict the behavior of an object, following the diagnosis of its present state, and select the control actions for prevention of failure. Figures 2; references: 10 Russian.
[238-2415]

UDC 681.322

SPECIAL-PURPOSE HYBRID MICROCOMPUTER FOR RAPID DIFFERENTIAL MEDICAL DIAGNOSIS

Kiev ELEKTRONNOYE MODELIROVANIYE in Russian Vol 5, No 3, May-Jun 83 pp 100-101

BARDACHENKO, VITALIY FEODOS'YEVICH, candidate of technical sciences, head Computation Center, Oblast Administration, Khmel'nitskiy;
NESTER, VLADIMIR VOYTSEKHOVICH, head of laboratory, Kiev

[Abstract] A special-purpose hybrid microcomputer for medical diagnosis has been developed jointly by the Institute of Modeling of Power Engineering Problems, UkSSR Academy of Sciences (Kiev), the "Kation" Industrial Association (Khmel'nitskiy) and the Chernovtsy Institute of Medicine. The computer can quantitatively estimate the diagnostic significance of a definite set of characteristic symptoms of various diseases in order to determine the chance of persons undergoing examination actually having these diseases. The computer has been designed for rapid diagnosis by the differential method. The microprocessor contains two timer-scalers which calculate the scalar product of a sum of resistances in the memory array by a scale factor proportional to the capacitance of the time setting RC circuit at the input of each. The medical memory is stored in a differential-diagnostic table with diseases listed in rows and symptoms listed in columns. The symptoms can be simulated numerically within 0.05% accuracy, the error of the microcomputer does not exceed 0.1%. Input data are processed in the hybrid mode and the results are displayed digitally. The microcomputer has been tested at the Chernovtsy Institute of Medicine as well as at the Vinnitsa Institute of Medicine and the Vinnitsa hospital imeni N. I. Pirogov. It has been recommended for production. The experimental prototype was displayed at the "Kardiologiya-82" Exhibition in Moscow and at a medical engineering exhibition for CEMA countries.
[238-2415]

UDC 621.373.078

COMPUTER MODELING OF FREQUENCY SYNTHESIZERS WITH PULSE-PHASE AUTOMATIC-CONTROL LOOP

Moscow ELEKTROSVYAZ' in Russian No 4, Apr 83
(manuscript received 17 Dec 80) pp 52-58

MALINOVSKIY, V. N. and ROMANOV, S. K.

[Abstract] Computer modeling of a digital frequency synthesizer for pulse-phase automatic frequency control is shown, which not only facilitates an analysis of processes in such a device without using closed-loop equations but also offers wide flexibility for structural modifications. Such a synthesizer is described accurately enough as an integral frequency-pulse modulator including two integrators, two relays, a controllable reference oscillator with low-pass filter, a variable-quotient frequency divider in the

feedback branch, and a comparator in the form of any pulse-phase detector or pulse-type frequency-phase detector. Each component is modeled with the aid of signal voltage-phase-time curves and binary tables. Processes in the model are calculated using transfer functions and matrix equations. A general algorithm of calculating the dynamic performance and the transient characteristics is constructed with sufficiently small time discretization steps. Theoretically this does not require more machine time than the algorithm of a mathematical model in the natural phase space. Figures 6; tables 5; references 10: 8 Russian, 2 Western (1 in translation).
[232-2415]

ELECTRON DEVICES

UDC: 621.382.2

INFLUENCE OF AMPLITUDE OSCILLATIONS ON NOISE IN AVALANCHE DIODE GENERATORS OF GALLIUM ARSENIDE WITH HOMOGENEOUS DOPING PROFILES

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 4 May 82) pp 380-388

KORNILOV, S. A. and PAVLOV, V. M., Leningrad Electrical Engineering Communications Institute

[Abstract] The purpose of this work was to produce experimental data which could fill gaps in existing experimental works which usually fail to present all the information necessary for calculations and which do not inspire confidience that the data published are typical. Studies were performed in the 10 GHz band on an avalanche diode of GaAs with a Schottky barrier and homogeneous doping profile. This type is attractive because it best satisfies the theoretical prerequisites of narrow breeding layer, equality of shock ionization coefficients and drift velocities of both electrons and holes. Three batches of similar diodes manufactured by identical technology at different times were studied. The results indicate that experimental testing of the theory of periodic unsteady fluctuations in such diodes should be performed at $I_0 < I_{nom}$ where the variation in level as a function of amplitude of oscillations is quite smooth. Figures 6; tables 5; references 13: 8 Russian, 5 Western.
[240-6508]

UDC: 621.385.6

AMPLITUDE FLUCTUATIONS OF BACKWARD WAVE TUBE IN SHORTWAVE PORTION OF MILLIMETER BAND

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 15 Oct 81; in final form 15 Sep 82) pp 370-374

GORONINA, K. A., Institute of Applied Physics, USSR Academy of Sciences

[Abstract] This work is concerned with an experimental study of the spectrum of amplitude fluctuations in a backward wave tube in the shortwave portion of the millimeter band at 0.15-2000 MHz, and is a continuation of an earlier work in which similar studies were performed for frequency fluctuations. The results of both works are discussed in order to explain the causes of the fluctuations. Measurements showed that the intensity of amplitude fluctuations, like that of frequency fluctuations, depends on the high voltage supply of the tube accelerating the electrons. There are a number of intensity minima and maxima in the generation range, their locations related to the so-called sectioning of the BWT characteristics generated by parasitic reflections. It is concluded that: 1) In accordance with the experimental observations the reasons for the observed frequency fluctuations are fluctuations in electron velocity equivalent to fluctuations in voltage; 2) The amplitude fluctuations are explained by the same factor as the frequency fluctuations; 3) Where dA/dV_y is near 0, the contribution of amplitude fluctuations resulting from velocity fluctuations is flight and additional instability factors appear; and 4) At 0 steepness, the value of $S_{\alpha\nu}$ is not equal to 0, indicating the existence of frequency fluctuations generated by the same additional factors which explain the amplitude fluctuations. Figures 4; references: 4 Russian.
[240-6508]

UDC: [66.013:658.26].003.1

ORGANIZATION OF ENERGY ACCOUNTING AT CHEMICAL INDUSTRY ENTERPRISES

Moscow PROMYSHLENNAYA ENERGETIKA in Russian No 5, May 83 pp 2-4

RYABTSEV, N. I., candidate of technical sciences, "Soyuzkhimpromenergo" (State Chemical Production Power) Production Association and
VYATKIN, M. A., candidate of economic sciences, All-Union Correspondence Polytechnical Institute

[Abstract] A discussion is presented of the economic desirability of installing electric meters throughout a chemical industry enterprise, so that the enterprise can determine the amount of electric power being consumed by the various processes underway. In each case, the desirability of installing a meter must be determined by balancing the potential energy savings against the cost of installing and operating the meter itself. Calculations can be used in order to determine the optimal number of meters and their optimal locations, so that in combination with calculations of the energy used based on known equipment characteristics, the meters can achieve precise accounting of energy utilization throughout the enterprise, minimizing energy use as a cost factor in the production of the end product of the enterprise. Neither total reliance on calculations based on published energy characteristics nor indiscriminant installation of electric meters is desirable as a means of minimizing energy costs. References: 4 Russian.
[245-6508]

PLANT POWER ENGINEER'S HANDBOOK: MERCURY SEALED-CONTACT REED RELAYS

Moscow PROMYSHLENNAYA ENERGETIKA in Russian No 5, May 83 pp 62-63

[Abstract] The USSR developed the design and mastered the production of locking and switching mercury sealed-contact reed relays. The relays are intended for switching ac and dc circuits in industrial automation devices, instrument making, computer technology, radioelectronics, measurement and pulse technology devices, electronic communications, illumination engineering, flash photometry, etc. The MKSR-45181 device is diagrammed. It is a sealed glass tube, at the ends of which are soldered leads of a magnetically soft material. When an external magnetic field is applied, a strip of metal wet with mercury alternates its position between the two leads which are at one end of the glass tube, effectively switching the contact path. These devices can control processes under the control of an electromagnetic or permanent magnet. The technical characteristics of the device include: closed contact resistance not over 0.03 ohms, open contact resistance at least $1 \cdot 10^{11}$ ohms, capacitance of open contact not over 0.5 pf, power cpacity 80 W without spark damping circuit, 250 W with spark damping circuit, current switched not over 3 A with active load, 5 A with inductive load, voltage switched not over 2500 V upon closure, 550 V upon opening, number of switching operations during service life at least $5 \cdot 10^9$, mass 6 g. [245-6508]

UDC: 538.56

NATURAL ELECTROMAGNETIC FIELD COORDINATES

Moscow IZVESTIYA AKADEMII NAUK SSSR: ENERGETIKA I TRANSPORT in Russian
No 2, Mar-Apr 83 (manuscript received 10 Feb 81; revised 31 May 82)
pp 87-94

GRACH, I. M., Frunze

[Abstract] The task of this work is to develop natural coordinates of an electromagnetic field, determine their properties and generate expressions defining the basic parameters of such a field in its coordinates. The natural coordinates of the electromagnetic field refer to a generalized curved system of coordinates x_1, x_2, x_3 with scale coefficients h_1, h_2, h_3 in which the electric field intensity vector E, the magnetic field intensity vector H and the Pointing vector Π have but one component each, tangential to the corresponding coordinate lines. The properties of the natural coordinates of the transverse electromagnetic field are defined. Expressions are derived for the integral and differential characteristics of a transverse electromagnetic field in tis natural coordinates. An example is appended in which it is required to find the expression for capacitance, inductance and electric and magnetic field intensities of a torroidal condensor with a circular cross section electrode with displaced axes. Figures 1; tables 2; references 8: 6 Russian, 2 Western in translation.
[242-6508]

UDC: 621.313.333

CALCULATION OF MAGNETIC FIELD OF SYNCHRONOUS SALIENT-POLE MACHINE

Moscow IZVESTIYA AKADEMII NAUK SSSR: ENERGETIKA I TRANSPORT in Russian
No 2, Mar-Apr 83 (manuscript received 25 Sep 81) pp 79-86

AFANAS'YEV, A. A. and PUPIN, V. M., Cheboksary

[Abstract] A study of the specifics of numerical calculation of the magnetic field in a synchronous salient-pole machine is made by the method of magnetic

conductivity with separate consideration of the entire spectrum of conductances of the various tooth structures in the air gap: the teeth of the stator and rotor, the poles of the inductor, nonuniformities of the gap between the pole tips and the surroundings of the armature bore. The method can be used to calculate magnetic fields in other types of electric machines with various types of gaps and tooth formations such as asynchronous and reducer machines. The calculation is performed, with the assumption that the magnetic field in the air gap is plane parallel, saturation of the tooth zone has no influence on the configuration of the magnetic field in the air gap and the magnetic field in the gap is a field of mutual induction. The numerical method suggested for calculating the magnetic field is basically more accurate than known methods of harmonic conductances. The solution of the initial equations can be implemented by a combined method using numerical methods for continuing the solution based on a parameter and by Newtons method. A specific example is appended. Figures 3; references: 9 Russian. [242-6508]

UDC: 621.372.2.001.24

DESIGN OF STRIP TRANSMISSION LINES BY METHOD OF PARTIAL AREAS CONSIDERING SPECIFICS AT THE EDGE

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 17 May 82) pp 357-362

ZARGANO, G. F., SINYAVSKIY, G. P. and TKACHENKO, V. P., Rostov State University

[Abstract] A study is made of an asymmetrical rectangular shielded strip line with a finite thickness of the central conductor and piecewise homogeneous dielectric filler which can be looked upon as a base model for the design of other types of lines: quadratic coaxial, grooved, three-strip symmetrical, high-Q. The task is performed in the quasi-static approximation which yields good results in most cases. To simplify the problem, the authors limit themselves to analysis of a flat line symmetrical with respect to plane BD, which divides the line into two equivalent halves. The problem is thus reduced to solution of a two-dimensional Laplace equation for a scalar field potential with certain boundary conditions. The convergence of the algorithm suggested was tested in calculating the parameters of a symmetrical rectangular shielded and square coaxial line with homogeneous air filling. Consideration of the specifics of behavior of the field near edges in the internal conductor allows significant improvements of convergence of the method and increases the accuracy of calculation results. Figures 4; tables 3; references 7: 5 Russian, 2 Western.
[240-6508]

UDC 621.372.8

PERFORMANCE CHARACTERISTICS OF SEMICONDUCTOR-TYPE CUTOUT SWITCHES IN MINIATURE WAVEGUIDES

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received 28 Sep 82) pp 64-66

GERMANOV, V. A. and STEBLEVA, N. G.

[Abstract] The performance characteristics of a cutout switch consisting of an n-i-p-i-n diode at the center between two rectangular quarter-wave-length waveguide segments with resonance slots are described. Filling the waveguide segments but not the slots with a dielectric makes it possible to reduce their dimensions by a factor proportional to the square root of the dielectric constant. A theoretical analysis indicates and experimental studies have confirmed that the dielectric filler also appreciably reduces the decoupling action of a diode, typically from 20-25 dB to 13-15 dB when the dielectric constant is $\varepsilon = 10$. This effect can be suppressed by mounting the diode in a dumb-bell diaphragm, the dimensions of the latter determining the resonance frequency. The dielectric filler increases the bandwidth of such a switch, while shifting it to lower frequencies, with corresponding decrease of the insertion loss and the VSWR. The decoupling characteristics can be further improved by use of multislot switches, according to data on 3-slot and 5-slot switches operating in the 3-cm wave band. The authors thank V. I. Vol'man for helpful comments and suggestions. Figures 3; references: 2 Russian.
[226-2415]

UDC 621.372.832

MICROWAVE TWO-PORT INVERTER CIRCUITS ON MICROSTRIP LINES WITH SYMMETRIC TRANSFER CHARACTERISTIC

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83
(manuscript received after completion 11 Aug 82) pp 61-63

BROVCHENKO, S. P. and MALYSHEV, V. A.

[Abstract] Microwave two-port inverting transmission lines are described where a low characteristic impedance has been made physically realizable by splitting the microstrip line into two halves of equal electrical lengths $1/2\theta$ and inserting between them a capacitive Π-network ($1/2C_2 - C_1 - 1/2C_2$) or T-network ($2C_1 - C_2 - 2C_1$). Expressions for the characteristic impedance of each version are derived, letting $n = C_1/C_0 l$ and $m = C_2/C_0 l$ (C_0- equivalent capacitance of microstrip line per unit length, l- length of microstrip line, $\theta = \omega l/v_\phi$ electrical length of microstrip line, ω- angular frequency, v_ϕ- phase velocity). The transfer function of these circuits is symmetric with respect to frequency. Figures 4; references 5: 3 Russian, 2 Western (both in translation).
[226-2415]

POWER ENGINEERING

UDC: 621.313.333.(-17)

EXPLOSION-SAFE ASYNCHRONOUS MOTORS FOR COLD CLIMATE REGIONS

Moscow PROMYSHLENNAYA ENERGETIKA in Russian No 5, May 83 pp 9-10

VOLKOVA, V. A., engineer, ZBARSKIY, L. A., candidate of technical sciences, ZHABROV, M. G., DUMA, A. B., POSTERNAK, M. I., engineers, All-Union Scientific-Research Institute of Explosion Protected and Mine Electric Equipment

[Abstract] The KhL2-KhL5 explosion-safe asynchronous cold-climate motors have been developed at the All-Union Scientific-Research Institute of Explosion Protected and Mine Electric Equipment on the basis of the new VAO2 series of devices. Rated at 132-315 kW at nominal voltages of 380 and 660 V, these devices are designed for use in areas with cold climate. Technical characteristics are presented in a table. In developing the design for the new motors, particular attention was given to assurance of reliability when operated in cold climate regions, as well as standardization with other series of motors. A cross sectional diagram of one model is presented. Experimental models of the motors have passed testing in terms of resistance to climatic factors with particular attention given to monitoring the explosion safety parameters, bearing units, insulation of the body and insulation between turns of the stator winding, paint and varnish coatings. Figures 1; tables 1; references: 5 Russian.
[245-6508]

UDC: 621.313.333.027.5.001.4

POSSIBILITY OF OPERATION OF EXPLOSION-SAFE HIGH-VOLTAGE MOTORS WITH 'MONOLIT-2' INSULATION AFTER PARTIAL IMMERSION IN WATER

Moscow PROMYSHLENNAYA ENERGETIKA in Russian No 5, May 83 pp 13-14

ZBARSKIY, L. A., candidate of technical sciences, PORSHNEV, Yu. V., SOROKO, P. A., MAKAGON, V. A., engineers, All-Union Scientific-Research Institute of Explosion Protected and Mine Electric Equipment

[Abstract] The authors' institute has performed laboratory studies of the possibility of operating series VAO2 6kV motors with a "Monolit-2" stator winding after partial immersion in water without subsequent drying. After immersion to a level below the projecting end of the motor shaft for periods of not over 4 days, these machines can be used without drying after the following steps are taken: 1) Removal of water from inner surface of electric lead device, all current conducting parts and porcelain insulators, which must be dried with a rag; and 2) measurement of insulation resistance of stator winding relative to motor body each hour for 6 to 8 hours of operation. The resistance should be no less than 6 MOhm. Figures 1; references: 2 Russian.
[245-6508]

UDC [621.315.14:621.3.014]:001.24

CALCULATION OF LOSSES AND FORCES IN PHASE SHIELDED CURRENT CONDUCTORS

Moscow IZVESTIYA AKADEMII NAUK SSSR: ENERGETIKA I TRANSPORT in Russian No 2, Mar-Apr 83 (manuscript received 16 Dec 81) pp 72-78

CHAL'YAN, K. M., Baku

[Abstract] Results are presented from calculation of electromagnetic parameters of the current conductors of 200-1000 MW generators under two short circuit conditions. The short circuit calculation mode characterizing the greatest electrodynamic action on the current conductors is a three-phase short circuit at the terminals of the generator unit. However, there is interest in calculating short circuits with a short at the lower side of the transformer unit, considering that calculation for this case is necessary in order to estimate the electrodynamic stability of joints designed to attach the internal power plant supply transformers. Calculations were performed for current leads with various connection plans of shields and fixed geometric dimensions and currents in the buses. The shielding factor representing the primary electromagnetic parameter acting on the bus of a given lead is practically independent of the pattern of current change in the bus with time. The actual electrodynamic forces acting on the buses were calculated by a method used in planning practice. They proved to be significantly greater than the forces based on a mathematical model, a result of the artifically elevated approximate shielding

factor used in standard design practice. In determining the electrodynamic forces from force pulses and considering the fact that in current leads with continuous shields the electrodynamic forces increase gradually from the moment of a short, reaching a maximum after about 0.1 s. The difference in electrodynamic forces calculated by the accepted methods are found to be still greater. Figures 2; tables 4; references 6: 4 Russian, 2 Western.
[242-6508]

ANALYSIS OF LABOR COSTS IN OPERATING 500-750 kV ELECTRIC POWER TRANSMISSION LINES

Moscow ENERGETIK in Russian No 5, May 83 pp 23-25

KOROBKOV, N. M., engineer, "Long-Distance Electrical Transmission" Production Association

[Abstract] The operation of electric power transmission lines includes maintenance, major repair and work related to elimination of the results of fault situations. Maintenance and capital repair represents some 99% of all the labor expended by operating and repair personnel. The expenditure of labor has been greatly reduced because of mechanization of maintenance and repair operations. At the author's production association, some 35 to 40% of major repair work is performed with the lines switched off. For example, in 1980 planned power line down-time for repairs amounted to 2133 hours, or 33.46 hours per 100 km. Emergency down-time amounted to 1.13 hours per 100 km. The following methods have been developed for performance of work with the lines still in operation: replacement of defective insulators, replacement of supports, minor repairs of conductors and replacement of individual types of defective elements, which amount to 95% of all work performed with the power off. Figures 5.
[244-6508]

UDC: 535.2

TRANSMISSION OF PULSED LIGHT SIGNAL THROUGH ABSORBING MEDIUM WITH STRONG ANISOTROPIC SCATTERING

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 6 Jul 82) pp 300-309

DOLIN, L. S., Institute of Applied Physics, USSR Academy of Sciences

[Abstract] A radiation transfer equation is used in order to summarize the results of earlier works concerned with the effect of random large scale heterogeneities in the refractive index of a medium on the envelope of a pulsed signal passing through the medium, in order to cover the case in which the scattering medium has a notable absorptive capacity, and the pulsed signal is radiated by an aperture of finite dimensions, producing a pulsed wave beam. The variation of delay and length of the signal received as functions of path length and initial beam width is studied. Figures 5; references 22: 15 Russian, 7 Western (1 in translation).
[240-6508]

UDC: 621.391

SPECIFICS OF DIFFRACTION OF LIGHT ON REGULAR REFLECTIONS OF ACOUSTIC WAVE

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian
Vol 26, No 3, Mar 83 (manuscript received 5 Oct 81) pp 334-338

KLUDZIN, V. V., Leningrad Institute of Aviation Instrument Building

[Abstract] The specifics are studied of the acousto-optical interaction of multiple reflections of an acoustical wave separated by a fixed angle. The geometry studied is presented in Figure 1. The use of the method of multiple acoustical reflections increases the information capacity of the system not only by increasing the length of the sample processed, but also by expanding the transmission bandwidth of the device if the reflections are deterministically oriented at an angle to each other. It is found that the system described has frequency-time discrete dispersion. Experimental and calculated data agree well. The most efficient solution would be the use of an exciter system in which the weak acoustical coupling between the piezo radiator and sound track would be used, e.g., as a nonresonant excitation system. Figures 4; references 6: 4 Russian, 2 Western.
[240-6508]

UDC 537.525

CURRENT-VOLTAGE CHARACTERISTICS OF HIGH-CURRENT ARC DISCHARGE IN VAPORS OF ELECTRODE MATERIAL

Kiev IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOELEKTRONIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 18 Feb 82) pp 94-96

SHENDAKOV, A. I., KRIZHANOVSKIY, V. I. and KUZ'MICHEV, A. I.

[Abstract] The current-voltage characteristics of an arc discharge in vapors of the electrode materials were measured under conditions of high current (1-20 kA) and pulsing voltage drop. The measurements were made with a shaping line matched to a noninductive load, the shaped square pulses being of 400 µs duration. The arc was ignited by a low-pressure auxiliary pulse discharge in a plane-parallel gap between cathode and anode, each made of pure hard metal or alloy. The voltage drop was measured by the method of two instruments in series, with an initial atmosphere of pure hydrogen under a pressure of 1-10 Pa. The voltage drop was found to increase with the current, first sharply within the 2-5 kA range and then slightly above that. Five materials with widely different thermophysical properties were tried as anode material: coper, LS59-1 brass, Kh18N10T stainless steel, nickel, and molybdenum, with the cathode made of copper in each case. The results obtained with a 4 mm wide interelectrode gap and a pressure of 5 Pa reveal a strong correlation between the voltage drop and the properties of the anode material, specifically its melting point, as well as between the trend of the current-voltage characteristic and the rate of anode material erosion. Increasing the interelectrode gap up to 10 mm, with the pressure maintained at 5 Pa and with both electrodes made of the same material (copper), was found to increase the voltage drop appreciably. This is attributable to an increasing potential fall across the positive column and to an increasing potential fall across the positive column and to an increasing leakage of charged and neutral particles from the discharge region. Control experiments wth discharge in vacuum (10^{-3} - 10^{-2} Pa) and in mercury vapor (10^{-1} Pa) have revealed that the voltage drop is almost independent of the nature and the pressure of the initial gas in the gap. Figures 3; references: 5 Russian.
[225-2415]

UDC: 621.317.37

POLARIZATION AND SPECTRAL CHARACTERISTICS OF OPEN RESONATORS WITH INTERNAL HETEROGENEITIES

Gorkiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY: RADIOFIZIKA in Russian Vol 26, No 3, Mar 83 (manuscript received 29 Jun 82) pp 318-328

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[Abstract] The spectral and polarization characteristics of an open resonator with a flat isotropic dielectric heterogeneity of arbitrary thickness are studied. The task is to determine the resonant oscillations in the system, the field of which must be a standing wave in all three basic areas. A method is suggested for studying the spectral and polarization properties of the resonator for the cases when the flat layer is made of various materials. The method is used to study the properties of the resonator with an isotropic dielectric layer in detail. It is shown for the first time that in such a resonator the wave beam is elliptical with elliptical polarization in the transverse cross section, and all of its components are found. A dispersion equation is produced with which the dielectric permeability of the isotropic dielectric can be defined and the spectral characteristics of the open resonator described. Figures 6; tables 1; references 16: 14 Russian, 2 Western in translation.
[240-6508]

IMPROVING QUALITY OF RADIOELECTRONIC HOME EQUIPMENT THROUGH EXTENSIVE APPLICATION OF MICROELECTRONICS

Moscow RADIOTEKHNIKA in Russian No 4, Apr 83 pp 95-96

GAL'PERIN, Ye. I., Scientific and Technical Society of Radio Engineering, Electronics and Communications imeni A. S. Popov

[Abstract] The second all-Union scientific-technical conference on application of integrated circuits in radioelectronic home equipment was held at Minsk in 1982, with participation of 170 specialists from 40 Soviet design offices, scientific research institutes and higher educational institutions. They reported on the progress made since the previous conference held in 1979 at Simferopol, progress based on extensive application of microelectronics and computer microprocessor technology. These reports and discussions indicate that further improvements will require: 1) Automation and simplification of tuning and control; 2) Optimization of systems designed complexly to improve the quality of video images, sound tracks, radio reception, and recording; 3) development of automatic performance control and automatic adaptation to operating conditions; 4) Development of standard wireless modules for remote control and programming of equipment operation; 5) Development of portable medical diagnostic and testing instruments; and 6) establishment of functional compatibility with monitoring and

data preliminary-processing devices for health care. This will call for new design concepts, new methods of digital data processing, and new techniques of infradyne reception with wideband preselection. This will also involve adaptation of computer-aided design facilities and hybrid-technology manufacturing facilities.
[226-2415]

EIGHTH CONGRESS OF POWER ENGINEERING AND ELECTRIC ENGINEERING INDUSTRY SCIENTIFIC AND TECHNICAL SOCIETY

Moscow PROMYSHLENNAYA ENERGETIKA in Russian No 5, May 83 p 60

KOZAKOV, A. K.

[Abstract] The 8th Congress of the NTOEiEP, held in November 1982 at Leningrad, was preceded by a broad publicity campaign intended to mobilize the society for successful implementation of the resolutions of the 26th CPSU Congress and the items of the 11th Five Year Plan. The conference was opened by N. N. Kovalev, chairman of the Central Administration of the organization, corresponding member of the USSR Academy of Sciences, followed by speeches by V. N. Budenny, deputy minister of power engineering and electrification of the USSR, Yu. A. Nikitin, deputy minister of the electrical industry, and Academician I. A. Budzko. All of them highly rated the assistance of workers in the NTOEiEP in developing the national economy. In 1981 alone, 1325 billion kW·hr of electricity were produced, specific fuel consumption falling to 327.1 g/(kW·hr), with the installed electric power-plant capacity exceeding 277 million kW. The participants noted the following unsolved problems: 1) Insufficient rates of scientific and technical progress; 2) Insufficient experimental capacity of many leading institutes; 3) Low level of mechanization of labor at many enterprises; 4) Failure to meet plans for introduction of new equipment; and 5) Lag in the production of the most progressive types of products. The delegates pointed out non-utilized reserves and emphasized the significance of the society in the intensification of production and improvement of its effectiveness. The congress adopted a resolution which amounts to a program for the work of the society for the next 5 years. A new central administration was elected, as were delegates to the 6th All-Union Congress of Scientific and Technical Societies.
[245-6508]

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